Mullard Technical Communications

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A new adjustor for Mullard ferrite inductor cores

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Mullard Research Laboratories

UDC 621.318.424: 621.3.072.32: 621.3.076.22

The new adjustor developed for the Mullard R-series of ferrite inductor cores is described. The results of measurements to determine the inductance stability under dry-heat cycling, bump, and vibration test conditions are presented, and adjustment curve wow and setting accuracy are discussed.

INTRODUCTION

The Mullard Vinkor range of ferrite pot cores to BS4061: Range 1 (Ref. 1) has now been in use for about 10 years. The most exacting requirements for these components occur in carrier telephony applications, where ferrite pot-core inductors are used as elements in filter and equaliser circuits. To meet the frequency stability specifications of these circuits, it is necessary to achieve inductor stabilities of a few hundred ppm. Hence the adjustor mechanism of ferrite pot cores, which can be a major contributory cause of instability, has received a great deal of attention.

The design of adjustors becomes progressively more difficult as the size of the ferrite core central hole decreases because of the increasing influence of manufacturing tolerances. During mid-1967 it was proposed to introduce the R-series of inductor cores. These are cores based on a square module, and are designed for use on printed-wiring boards. They range in size from R6 to R10, the numeral indicating the size of the module in tenths of an inch. The R6 and R7 sizes have a 3·10/3·05mm central hole which is smaller than the hole in the equivalent Vinkor, for example 4·65/4·60mm in the 18mm Vinkor. Experience with the 18mm IEC core, which has a 3·10/3·05mm hole, has shown that satisfactory adjustor performance is difficult to achieve in a hole of this size. Therefore it was decided to prepare a new adjustor design specifically for the R-series cores. The object of the development was to achieve the 18mm Vinkor adjustor inductance stability of less than 500 ppm (0·05%) in a ferrite core with a 3·10/3·05mm central hole.

VINKOR ADJUSTOR DESIGN

The inductance adjustment in a pot core is obtained by varying the position of a ferrite tube which shunts the air gap, as shown in Fig. 1. The calculation of adjustment curves is now well known and is described elsewhere (Ref. 2). To achieve an adequate adjustment range when the air gap is small, it is necessary to have a relatively small clearance between the ferrite tube and the central hole. It therefore follows that the ferrite tube must move concentrically in the central hole to give a smooth adjustment curve.

The adjustor used with the 14 to 45mm Vinkors is shown in Fig. 2. A ferrite tube is attached to a plastic carrier which is screwed up or down on a threaded stud. The carrier has a round head with three equispaced vertical ribs which provide friction to hold the adjustor in position. This carrier is injection-moulded through the ferrite tube so that the outer surface of the tube is held concentric with the ribs on the head. The ferrite tube is retained by the shoulder at the bottom of the carrier. The central hole, in the lower part of the carrier, has three equispaced flats where the thread is impressed as the adjustor is screwed on to its stud. The stud has a fine thread, 12 or 14BA, which gives good adjustment accuracy. This arrangement gives a shake-proof adjustment that is free of backlash. The carrier is moulded from a high heat endurance grade of polypropylene, the short-term heat resistance rating being 150°C.

R-CORE ADJUSTOR DESIGN

The new adjustor for use with the R-series of ferrite cores currently being introduced is shown in Fig. 3. The design retains the main features of the Vinkor adjustment mechanism, in particular a polypropylene carrier having a central hole with three flats which engages with a finely threaded stud, 14BA on R6 and R7 cores.

The main design change concerns the head of the carrier and the way in which the friction between the adjustor carrier and the ferrite core central hole is provided. A skirt at the top of the carrier now replaces the three ribs

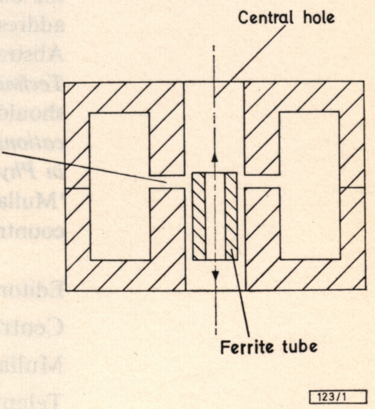

Fig. 1—Cross-section of ferrite pot core

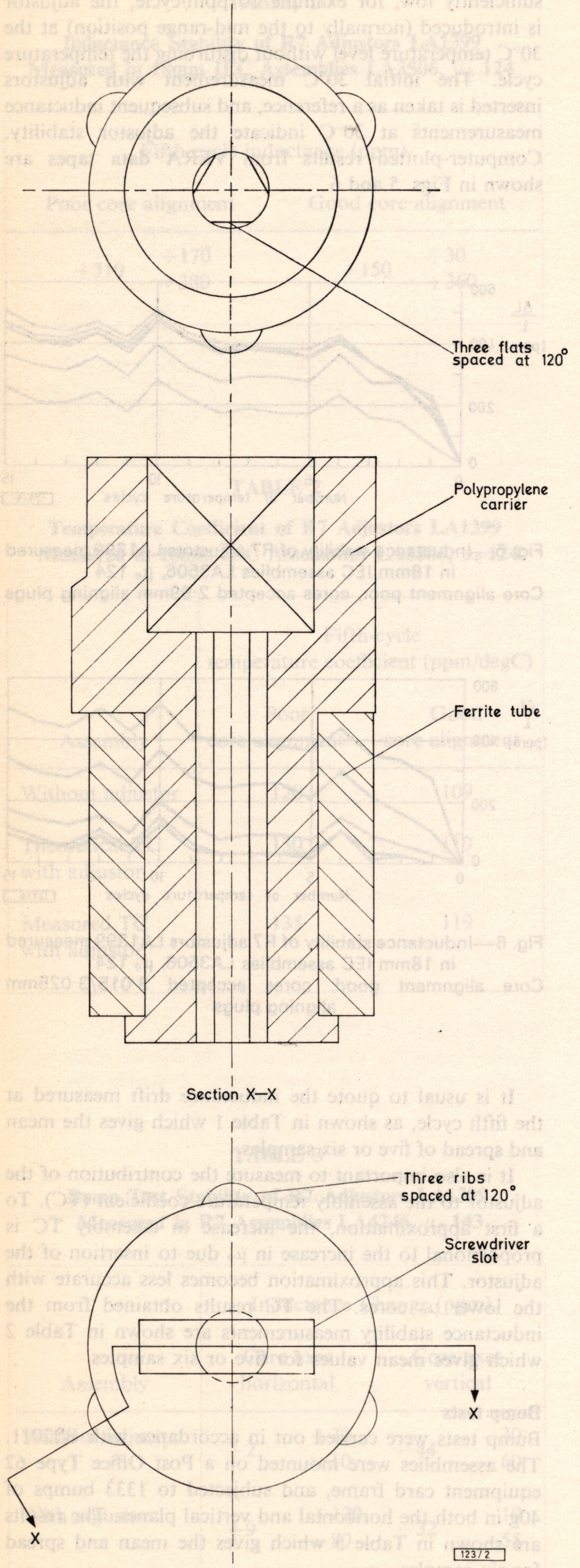

Fig. 2—18mm Vinkor adjustor for 4·65/4·60mm core hole

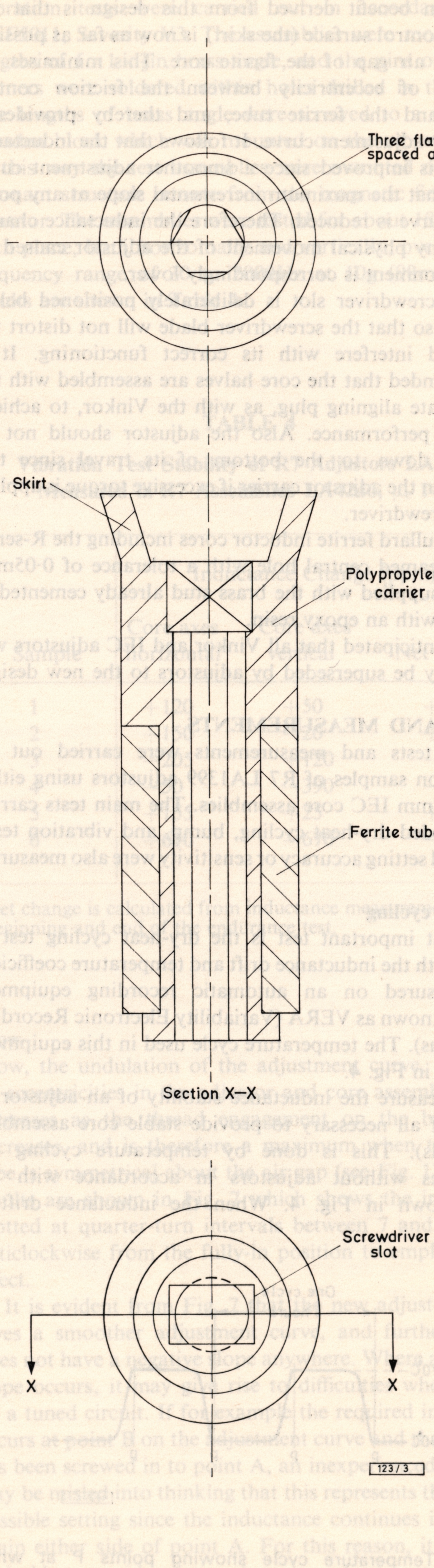

Fig. 3—R7 adjustor for 3·10/3·05mm core hole

immediately above the ferrite tube on the Vinkor adjustor. The main benefit derived from this design is that the friction control surface (the skirt) is now as far as possible from the air gap of the ferrite core. This minimises the influence of eccentricity between the friction control surface and the ferrite tube, and thereby provides a smoother adjustment curve. It follows that the inductance stability is improved since a smoother adjustment curve implies that the maximum incremental slope at any point on the curve is reduced. Therefore the inductance change due to any physical movement of the adjustor caused by the environment is correspondingly lower.

The screwdriver slot is deliberately positioned below the skirt so that the screwdriver blade will not distort the skirt and interfere with its correct functioning. It is recommended that the core halves are assembled with the appropriate aligning plug, as with the Vinkor, to achieve the best performance. Also the adjustor should not be screwed down to the bottom of its travel since this may strain the adjustor carrier if excessive torque is applied to the screwdriver.

All Mullard ferrite inductor cores including the R-series have a reamed central hole with a tolerance of 0·05mm, and are supplied with the brass stud already cemented in position with an epoxy resin.

It is anticipated that all Vinkor and IEC adjustors will eventually be superseded by adjustors to the new design.

TESTS AND MEASUREMENTS

All the tests and measurements were carried out on production samples of R7 LA1399 adjustors using either R7 or 18mm IEC core assemblies. The main tests carried out included dry-heat cycling, bump, and vibration tests. Wow and setting accuracy or sensitivity were also measured.

Dry-heat cycling

The most important test is the dry-heat cycling test in which both the inductance drift and temperature coefficient are measured on an automatic recording equipment (Ref. 3) known as VERA (Variability Electronic Recording Apparatus). The temperature cycle used in this equipment is shown in Fig. 4.

To measure the inductance stability of an adjustor, it is first of all necessary to provide stable core assemblies (test beds). This is done by temperature cycling the assemblies without adjustors in accordance with the cycle shown in Fig. 4. When the inductance drift is sufficiently low, for example 10 ppm/cycle, the adjustor is introduced (normally to the mid-range position) at the 30°C temperature level without disturbing the temperature cycle. The initial 30°C measurement with adjustors inserted is taken as a reference, and subsequent inductance measurements at 30°C indicate the adjustor stability. Computer-plotted results from VERA data tapes are shown in Figs. 5 and 6.

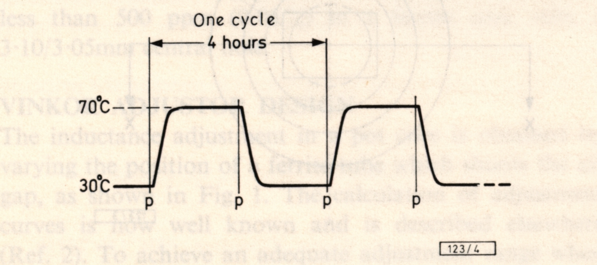

Fig. 4—Temperature cycle showing points P at which inductance is measured

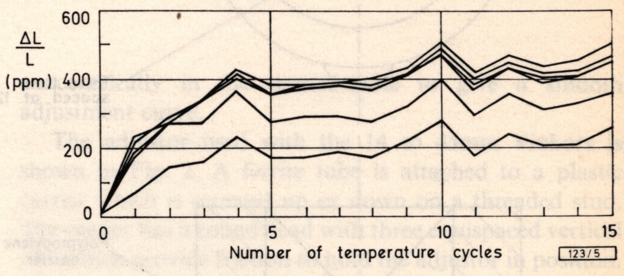

Fig. 5—Inductance stability of R7 adjustors LA1399 measured in 18mm IEC assemblies LA3506, μ_e 124
Core alignment poor, cores accepted 2·99mm aligning plugs

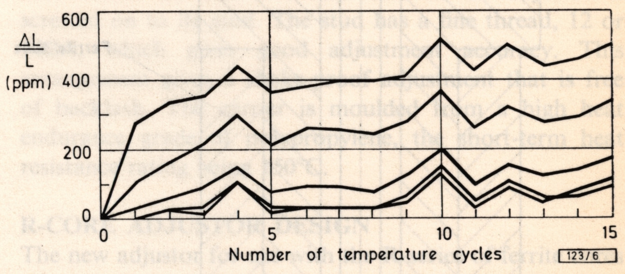

Fig. 6—Inductance stability of R7 adjustors LA1399 measured in 18mm IEC assemblies LA3506, μ_e 124
Core alignment good, cores accepted 3·015/3·025mm aligning plugs

It is usual to quote the inductance drift measured at the fifth cycle, as shown in Table 1 which gives the mean and spread of five or six samples.

It is also important to measure the contribution of the adjustor to the assembly temperature coefficient (TC). To a first approximation, the increase in assembly TC is proportional to the increase in μ_e due to insertion of the adjustor. This approximation becomes less accurate with the lower μ_e cores. The TC results obtained from the inductance stability measurements are shown in Table 2 which gives mean values for five or six samples.

Bump tests

Bump tests were carried out in accordance with BS2011. The assemblies were mounted on a Post Office Type 62 equipment card frame, and subjected to 1333 bumps of 40g in both the horizontal and vertical planes. The results are shown in Table 3 which gives the mean and spread for six samples.

TABLE 1

Inductance Stability of R7 Adjustors LA1399 Measured in 18mm IEC Assemblies LA3506, μ_e 124

Fifth-cycle inductance (ppm)			
Poor core alignment		Good core alignment	
+310	+170 +380	+150	+30 +360

TABLE 2

Temperature Coefficient of R7 Adjustors LA1399 Measured in 18mm IEC Assemblies LA3506, μ_e 124

Assembly	Fifth-cycle temperature coefficient (ppm/degC) Poor core alignment	Good core alignment
Without adjustor	120	109
Theoretical TC with adjustor	130	119
Measured TC with adjustor	135	119

TABLE 3

Bump Test Stability of R7 Adjustors LA1399 Measured in R7 Assemblies LA4246, μ_e 143

Assembly	Inductance Change (ppm) Core axes horizontal		Core axes vertical	
Without adjustor	+2	+30 −10	−44	−30 −60
With adjustor	+9	+120 −80	−32	−10 −55

Vibration tests

Vibration tests were carried out in accordance with DEF5011, Severity V3. The assemblies were mounted on lengths of $1 \times 1 \times \frac{1}{8}$in brass angle, and the pins of the coil formers were soldered within holes drilled in the angle. The lengths of brass angle were screwed to a 6in thick wooden block which was mounted on the vibrator table. In this way the test assemblies were mounted on a non-resonant structure remote from the magnetic field of the vibrator. The assemblies were tested for about 18 hours in each plane, both horizontally and vertically, over a swept frequency range of 30 to 2000Hz at 10g (98m/s^2). The results are shown in Table 4.

TABLE 4

Vibration Test Stability of R7 Adjustors LA1399 Measured in R7 Assemblies LA4246, μ_e 143

Sample	Inductance Change (ppm) Core axes horizontal	Core axes vertical	Net change*
1	+120	+50	+170
2	+150	−20	+130
3	−205	+120	−85
4	−50	−390	−440
5	+155	+25	+180
6	+610	−670	−60

*Net change is calculated from inductance measurements at the beginning and end of the endurance test.

Wow

Wow, the undulation of the adjustment curve, is caused by eccentricities in the adjustor and core assembly. Wow increases as the thread engagement on the brass stud decreases, and is therefore a maximum when the ferrite tube is symmetrical about the air gap (see Fig. 1). Typical results are shown in Fig. 7 which shows the inductance plotted at quarter-turn intervals between 7 and 10 turns anticlockwise from the fully-in position to emphasise the effect.

It is evident from Fig. 7 that the new adjustor design gives a smoother adjustment curve, and furthermore it does not have a negative slope anywhere. Where a negative slope occurs, it may give rise to difficulties when setting up a tuned circuit. If for example the required inductance occurs at point B on the adjustment curve and the adjustor has been screwed in to point A, an inexperienced operator may be misled into thinking that this represents the nearest possible setting since the inductance continues increasing again either side of point A. For this reason, it is highly desirable to avoid negative slopes on adjustment curves.

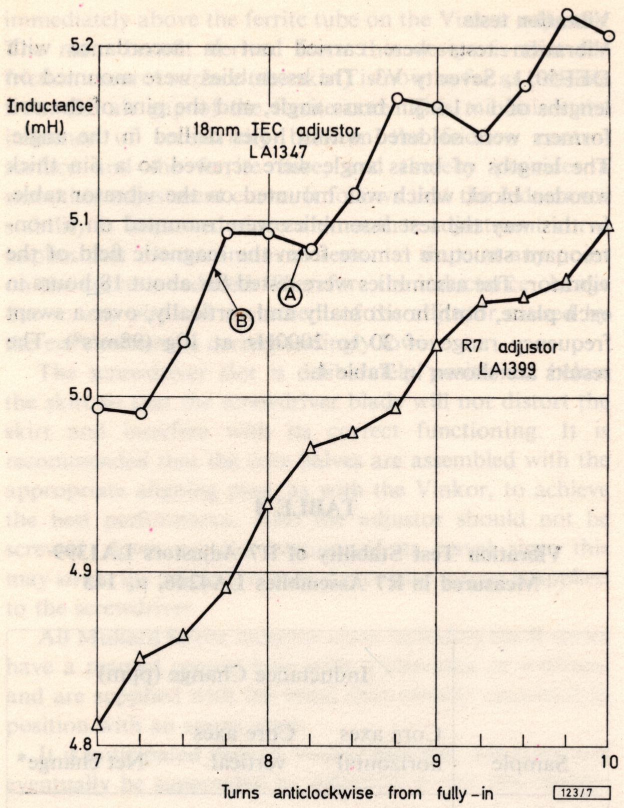

Fig. 7—Wow measurements using 18mm IEC assembly LA3507, μ_e 79·3, which accepted 3·005mm aligning plug

Setting accuracy

The setting accuracy of the R7 adjustor LA1399 was demonstrated on six LA4246, μ_e 143 assemblies using a free-running oscillator. Each inductor was connected in turn to the oscillator and resonated at about 95kHz. From the individual frequency measurements, a new frequency was calculated corresponding to the mid-range position of each adjustor. The adjustors were then inserted and adjusted as closely as possible to the required frequency. Table 5 shows the required and actual frequency settings achieved.

It can be seen from Table 5 that the setting accuracy is better than ±0·02% or ±200 ppm. The adjustment curve for this particular adjustor and assembly has a mid-range slope of 2·68%/turn, which means that the inductance change for 1 degree of rotation is 0·0075% or 75 ppm. Therefore to achieve a setting accuracy of ±0·02%, the adjustor must be set within about ±3 degrees of the exact position.

TABLE 5

Setting Accuracy of R7 Adjustors LA 1399
Measured in R7 Assemblies LA4246, μ_e 143

Sample	Required frequency (Hz)	Actual frequency (Hz)	Frequency error (Hz)	Frequency error (ppm)	Inductance error (ppm)
1	91682	91686	4	45	90
2	91461	91468	7	75	150
3	91657	91655	2	20	40
4	91691	91699	8	85	170
5	91985	91982	3	35	70
6	91517	91517	0	0	0

TABLE 6

R-core Adjustors which may be used in Vinkors or IEC Pot Cores

R-core		Vinkor	IEC pot core
R10	LA1428 LA1432 LA1433	45, 35, 30, 25mm	36, 30, 26mm
R7	LA1399 LA1427 LA1400	—	18mm

INTERCHANGEABILITY

It will be possible to use some of the R-core adjustors in Vinkors or IEC pot cores. Table 6 shows the interchangeability.

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High-quality audio amplifiers

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UDC 621.375.029.45

Three high-quality audio amplifiers, using silicon transistors, are described. The amplifiers supply output powers of 15W, 25W and 35W into an 8Ω load.

The design of these amplifiers is based on a complement of silicon transistors which, with the exception of the output pair, are all plastic-encapsulated devices. The output devices are matched hometaxial transistors encapsulated in metal TO-3 envelopes and designed for high power dissipation at high supply voltages. A quasi-complementary, single-ended class B push-pull configuration is used. The n-p-n output transistors are driven by a complementary pair which are preceded by a class A stage and a pre-amplifier stage. Another transistor provides stabilisation of the quiescent conditions.

The basic design is described and component values given for the application of the design to produce amplifiers with output powers of 15, 25, and 35W into an 8Ω load, with typically 0.1% total harmonic distortion at a frequency of 1kHz. Subdivisions in each section give details of individual amplifiers; circuit diagrams are given in Figs. 1, 2 and 3.

The amplifiers withstand normal overdrive conditions, but short-circuit protection, differing in detail for each amplifier, is recommended.

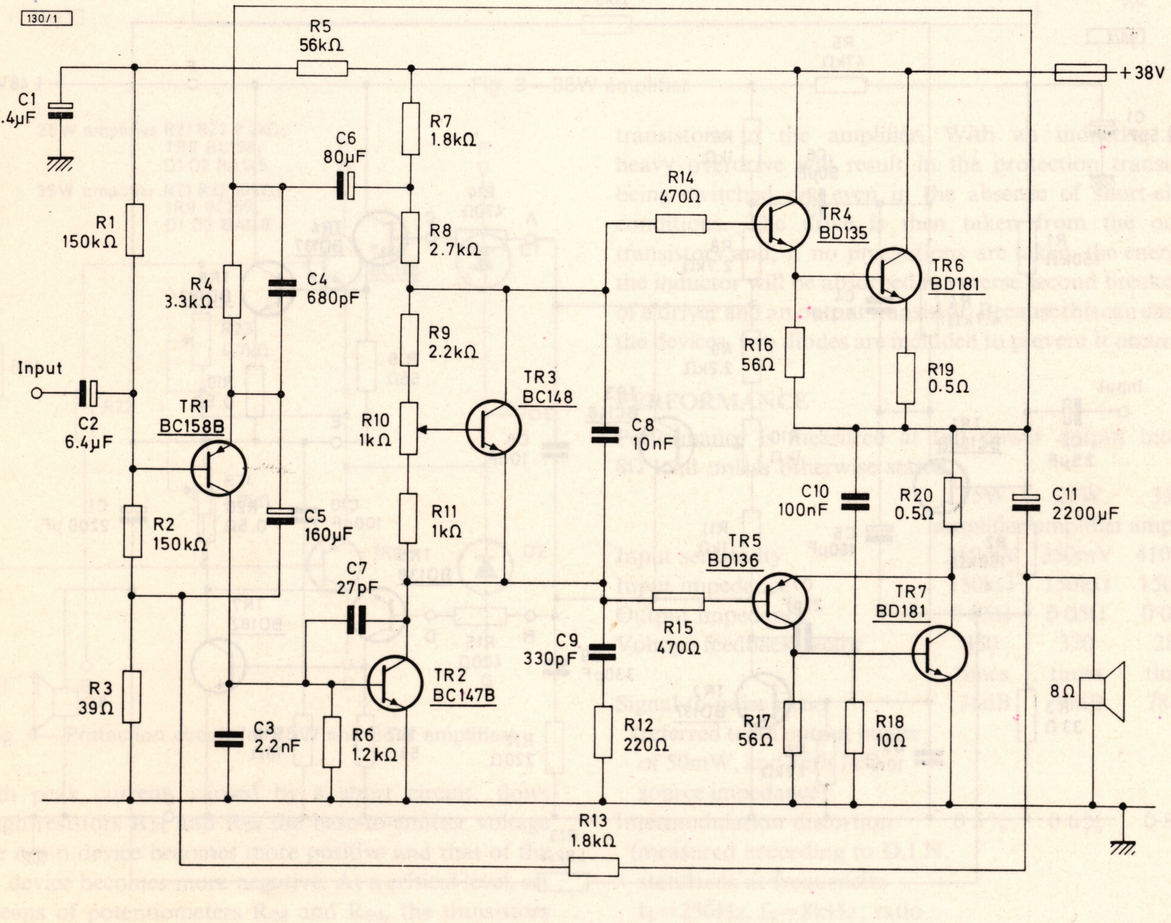


Fig. 1—15W amplifier

CIRCUIT DESCRIPTION

Pre-amplifier stage TR$_1$

Transistor TR$_1$ functions as a pre-amplifier and mid-point voltage stabiliser. A high-gain transistor is used to allow high values of a.c. and d.c. feedback to be applied via resistors R$_3$, R$_4$ and R$_{13}$.

Pre-driver stage TR$_2$

A class A amplifier stage supplies a drive current to the complementary drivers, TR$_4$ and TR$_5$. A low-value capacitor C$_7$ is connected between collector and base to limit the high-frequency gain of the stage. Resistors R$_7$ and R$_8$, and capacitor C$_6$, constitute a bootstrapping network.

Quiescent current stabilisation TR$_3$

Transistor TR$_3$ compensates variations in quiescent current caused by ambient temperature and supply voltage variations. The quiescent current is set by adjustment of preset potentiometer R$_{10}$.

Complementary driver and output stages TR$_4$ to TR$_7$

The matched driver transistors function as a phase inverter, and supply drive currents to the output pair and current losses in base-to-emitter resistors R$_{16}$ and R$_{17}$. To limit the maximum drive current for the driver and output transistors in the event of a short circuit of the loudspeaker terminals, 470Ω resistors are connected in series with the bases of the driver transistors. The heatsink requirements for transistors in these stages are discussed in the "*Components*" section.

D.C. Conditions

The d.c. conditions for the amplifiers are tabulated below.

	15W amplifier	25W amplifier	35W amplifier
Nominal supply voltage	38V	48V	60V
Collector current of TR$_1$	500μA	500μA	400μA
Collector current of TR$_2$	4mA	4mA	7·7mA
Quiescent current in driver pair	10mA	10mA	10mA
Quiescent current in output pair	40mA	40mA	40mA
Total current consumption (full output, at 1kHz)	625mA	830mA	1·1A

SHORT-CIRCUIT PROTECTION

15W amplifier

Adequate short-circuit protection is provided by an 800mA fast fuse connected into the supply lead.

25W and 35W amplifiers

Fig. 4 shows a method of protecting these amplifiers, the letters referring to points of connection in the amplifier circuits. The transistors are normally switched off. When a high peak current, caused by a short circuit, flows through resistors R$_{21}$ and R$_{22}$ the base-to-emitter voltage of the n-p-n device becomes more positive and that of the p-n-p device becomes more negative. At a critical level, set by means of potentiometers R$_{23}$ and R$_{24}$, the transistors are switched on so that current is drained from the driver transistors in the amplifier. With an inductive load, heavy overdrive will result in the protection transistors being switched on, even in the absence of short-circuit conditions. The drive is then taken from the output transistors and, if no precautions are taken, the energy in the inductor will be absorbed in reverse second breakdown of a driver and an output transistor. Because this can damage the devices, two diodes are included to prevent it occurring.

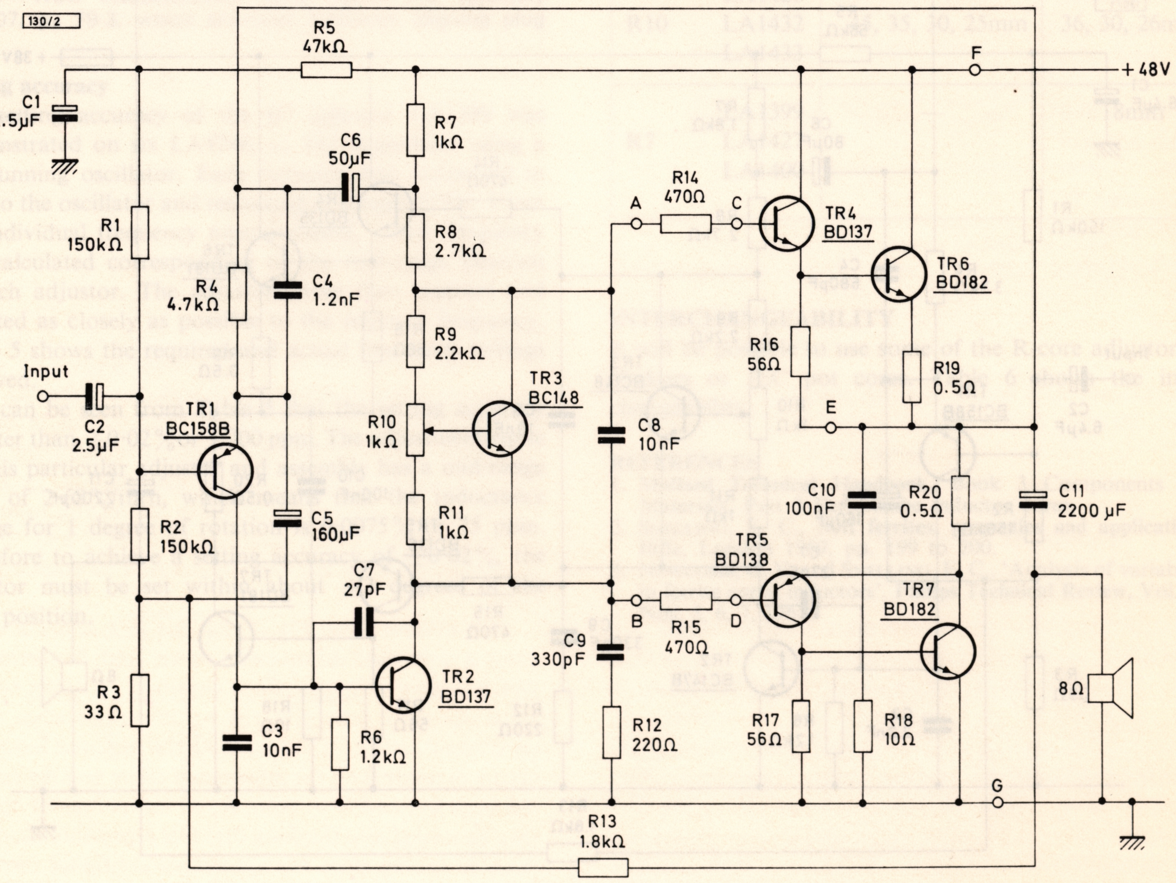

130/2
R5 47kΩ
F
+48V
C1 2.5μF
R7 1kΩ
C6 50μF
R1 150kΩ
R8 2.7kΩ
R4 4.7kΩ
C4 1.2nF
R9 2.2kΩ
A
R14 470Ω
C
TR4 BD137
TR6 BD182
Input
C2 2.5μF
TR1 BC158B
R10 1kΩ
TR3 BC148
C8 10nF
R16 56Ω
R19 0.5Ω
E
R2 150kΩ
C5 160μF
R11 1kΩ
C10 100nF
R20 0.5Ω
C11 2200μF
C7 27pF
TR5 BD138
B
R15 470Ω
D
TR7 BD182
C9 330pF
TR2 BD137
R3 33Ω
C3 10nF
R6 1.2kΩ
R12 220Ω
R17 56Ω
R18 10Ω
8Ω
G
R13 1.8kΩ

Fig. 2—25W amplifier

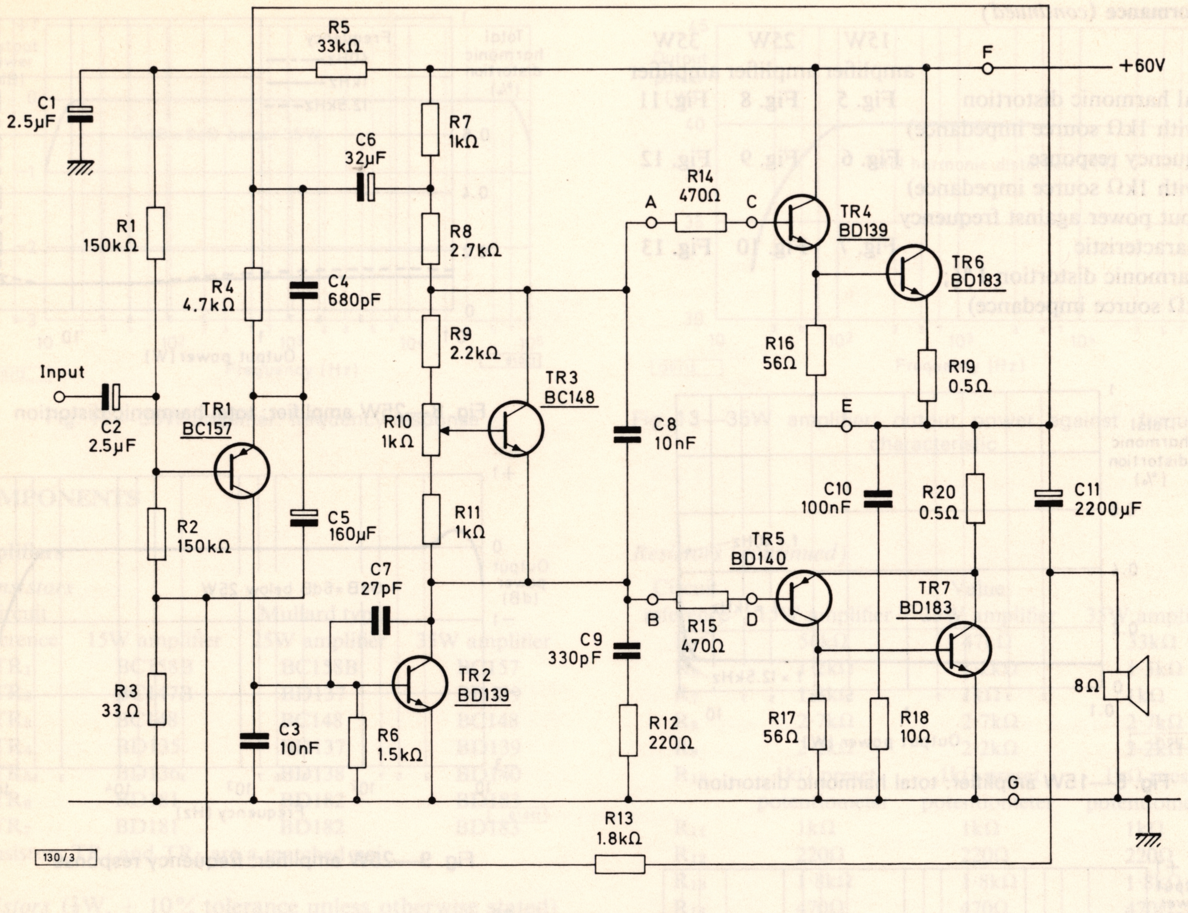
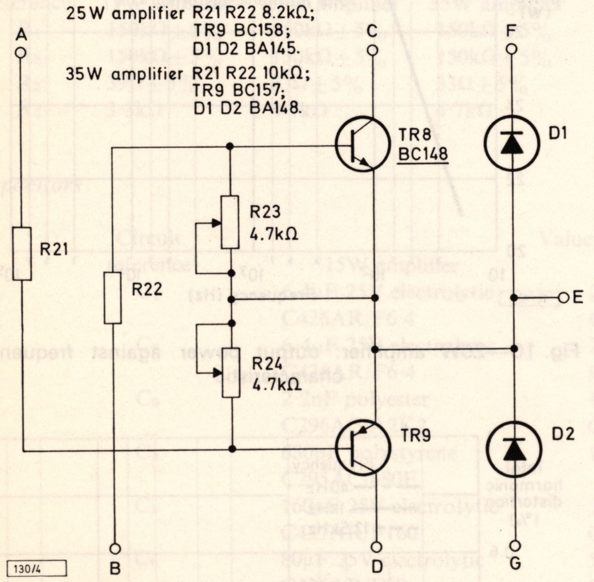

R5 33kΩ
F
+60V
C1 2.5μF
R7 1kΩ
C6 32μF
R14 470Ω
A
C
TR4 BD139
R1 150kΩ
R8 2.7kΩ
TR6 BD183
R4 4.7kΩ
C4 680pF
R16 56Ω
R9 2.2kΩ
R19 0.5Ω
Input
TR3 BC148
TR1 BC157
R10 1kΩ
C8 10nF
E
C2 2.5μF
C10 100nF
R20 0.5Ω
C11 2200μF
C5 160μF
R11 1kΩ
R2 150kΩ
TR5 BD140
C7 27pF
TR7 BD183
B
R15 470Ω
D
C9 330pF
TR2 BD139
8Ω
R3 33Ω
C3 10nF
R6 1.5kΩ
R12 220Ω
R17 56Ω
R18 10Ω
G
130/3
R13 1.8kΩ

Fig. 3—35W amplifier


A
C
F
TR8 BC148
D1
R21
R23 4.7kΩ
R22
E
R24 4.7kΩ
TR9
D2
130/4
B
D
G

25W amplifier R21 R22 8.2kΩ;
TR9 BC158;
D1 D2 BA145.

35W amplifier R21 R22 10kΩ;
TR9 BC157;
D1 D2 BA148.

Fig. 4—Protection circuit for 25W and 35W amplifiers

PERFORMANCE

Performance is measured at full power output into an 8Ω load unless otherwise stated.

	15W amplifier	25W amplifier	35W amplifier
Input sensitivity	350mV	350mV	410mV
Input impedance	150kΩ	150kΩ	150kΩ
Output impedance	0·05Ω	0·05Ω	0·05Ω
Voltage feedback factor	350 times	370 times	280 times
Signal-to-noise ratio (referred to an output power of 50mW, and with 1kΩ source impedance)	76dB	76dB	78dB
Intermodulation distortion (measured according to D.I.N. standards at frequencies f$_1$=250Hz, f$_2$=8kHz; ratio of signal voltages at f$_1$: f$_2$=4:1)	0·5%	0·6%	0·8%

Performance (*continued*)

	15W amplifier	25W amplifier	35W amplifier
Total harmonic distortion (with 1kΩ source impedance)	Fig. 5	Fig. 8	Fig. 11
Frequency response (with 1kΩ source impedance)	Fig. 6	Fig. 9	Fig. 12
Output power against frequency characteristic (harmonic distortion 1%; 1kΩ source impedance)	Fig. 7	Fig. 10	Fig. 13

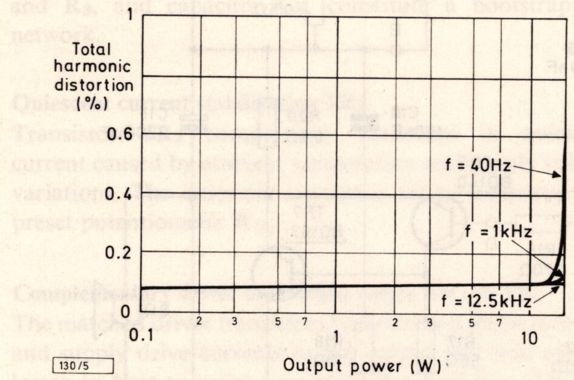


Fig. 5—15W amplifier: total harmonic distortion

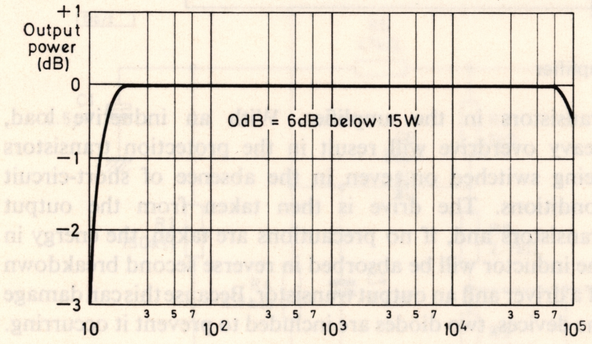


Fig. 6—15W amplifier: frequency response

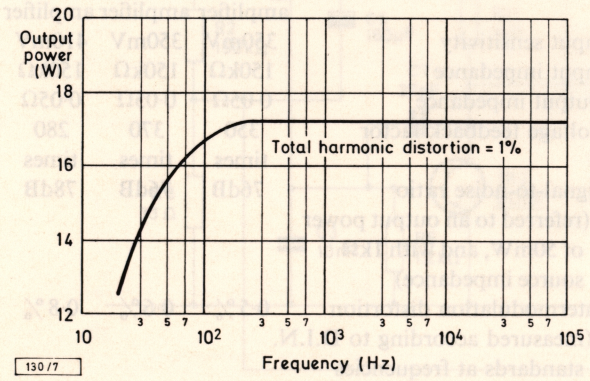


Fig. 7—15W amplifier: output power against frequency characteristic

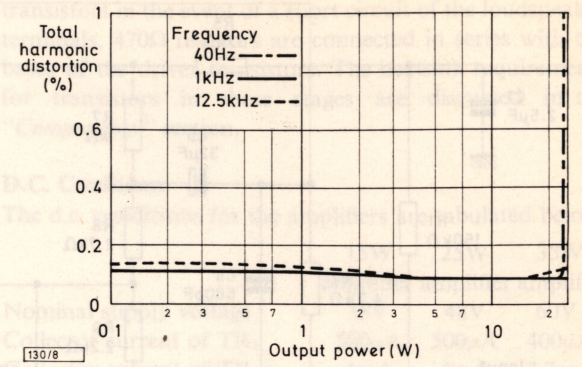


Fig. 8—25W amplifier: total harmonic distortion

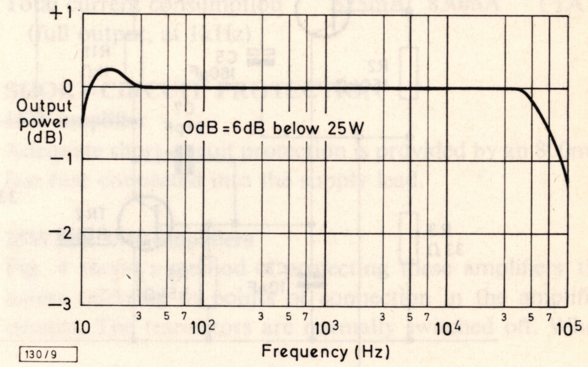


Fig. 9—25W amplifier: frequency response

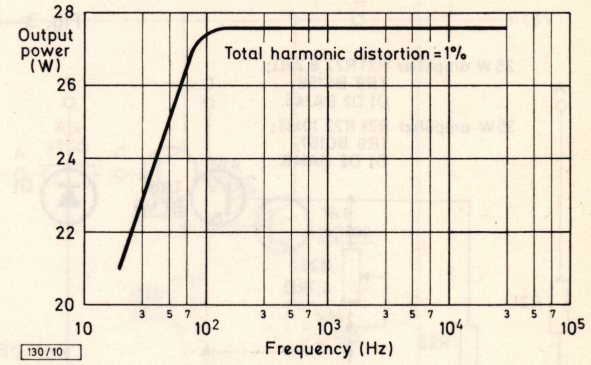


Fig. 10—25W amplifier: output power against frequency characteristic

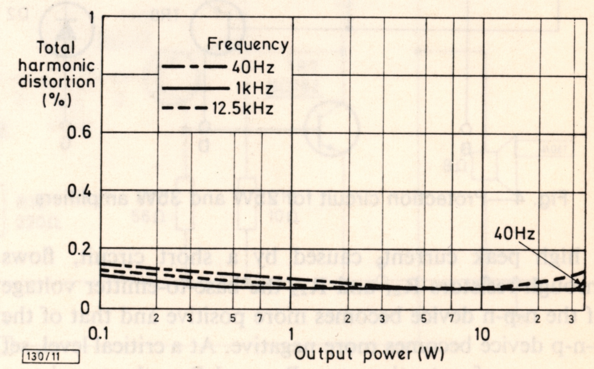


Fig. 11—35W amplifier: total harmonic distortion

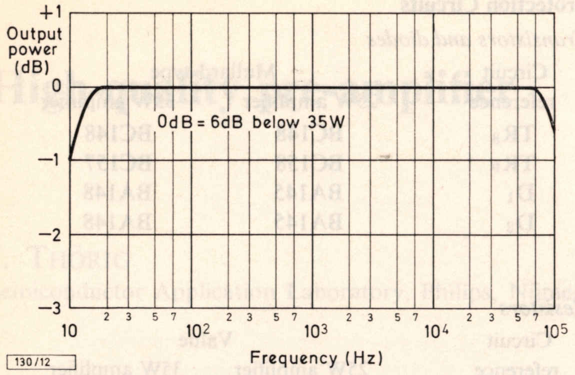


Fig. 12—35W amplifier: frequency response

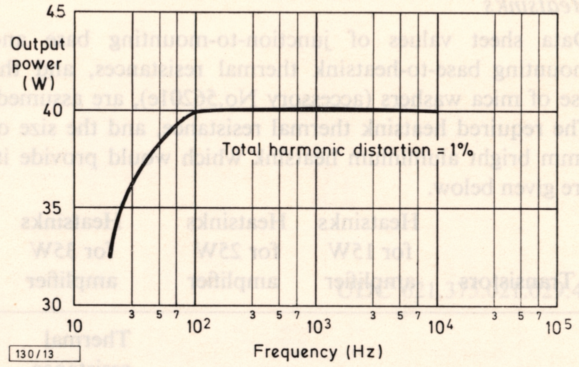


Fig. 13—35W amplifier: output power against frequency characteristic

COMPONENTS

Amplifiers

Transistors

Circuit reference	Mullard type 15W amplifier	25W amplifier	35W amplifier
TR_1	BC158B	BC158B	BC157
TR_2	BC147B	BD137	BD139
TR_3	BC148	BC148	BC148
TR_4	BD135	BD137	BD139
TR_5	BD136	BD138	BD140
TR_6	BD181	BD182	BD183
TR_7	BD181	BD182	BD183

Transistors TR_4 and TR_5 are a matched pair.

Resistors (⅛W, ± 10% tolerance unless otherwise stated).

Circuit reference	Value 15W amplifier	25W amplifier	35W amplifier
R_1	150kΩ±5%	150kΩ±5%	150kΩ±5%
R_2	150kΩ±5%	150kΩ±5%	150kΩ±5%
R_3	39Ω±5%	33Ω±5%	33Ω±5%
R_4	3·3kΩ	4·7kΩ	4·7kΩ

Resistors (continued)

Circuit reference	Value 15W amplifier	25W amplifier	35W amplifier
R_5	56kΩ	47kΩ	33kΩ
R_6	1·2kΩ	1·2kΩ	1·5kΩ
R_7	1·8kΩ	1kΩ	1kΩ
R_8	2·7kΩ	2·7kΩ	2·7kΩ
R_9	2·2kΩ	2·2kΩ	2·2kΩ
R_{10}	1kΩ preset potentiometer	1kΩ preset potentiometer	1kΩ preset potentiometer
R_{11}	1kΩ	1kΩ	1kΩ
R_{12}	220Ω	220Ω	220Ω
R_{13}	1·8kΩ	1·8kΩ	1·8kΩ
R_{14}	470Ω	470Ω	470Ω
R_{15}	470Ω	470Ω	470Ω
R_{16}	56Ω	56Ω	56Ω
R_{17}	56Ω	56Ω	56Ω
R_{18}	10Ω	10Ω	10Ω
R_{19}	0·5Ω, 2W	0·5Ω, 2W	0·5Ω, 5W
R_{20}	0·5Ω, 2W	0·5Ω, 2W	0·5Ω, 5W

Capacitors

Circuit reference	Value, description and Mullard type 15W amplifier	25W amplifier	35W amplifier
C_1	6·4μF 25V electrolytic C426AR/F6·4	2·5μF 64V electrolytic C426AR/H2·5	2·5μF 64V electrolytic C426AR/H2·5
C_2	6·4μF 25V electrolytic C426AR/F6·4	2·5μF 64V electrolytic C426AR/H2·5	2·5μF 64V electrolytic C426AR/H2·5
C_3	2·2nF polyester C296AC/A2K2	10nF metallised film C280AE/P10K	10nF metallised film C280AE/P10K
C_4	680pF polystyrene C295AC/680E	1·2nF polystyrene	680pF polystyrene C295AC/680E
C_5	160μF 25V electrolytic C437AR/F160	160μF 40V electrolytic C437AR/G160	160μF 40V electrolytic C437AR/G160
C_6	80μF 25V electrolytic C426AR/F80	50μF 25V electrolytic C426AR/F50	32μF 64V electrolytic C426AR/H32
C_7	27pF ceramic plate C333CC/C27E	27pF ceramic plate C333CC/C27E	27pF ceramic plate C333CC/C27E
C_8	10nF metallised film C280AE/P10K	10nF metallised film C280AE/P10K	10nF metallised film C280AE/P10K
C_9	330pF polystyrene	330pF polystyrene	330pF polystyrene
C_{10}	100nF metallised film C280AE/P100K	100nF metallised film C280AE/P100K	100nF metallised film C280AE/P100K
C_{11}	2200μF 25V electrolytic 071 16222	2200μF 40V electrolytic 071 17222	2200μF 63V electrolytic 071 18222

Heatsinks

Data sheet values of junction-to-mounting base and mounting base-to-heatsink thermal resistances, and the use of mica washers (accessory No.56201e), are assumed. The required heatsink thermal resistance, and the size of 2mm bright aluminium heatsink which would provide it, are given below.

Transistors	Heatsinks for 15W amplifier	Heatsinks for 25W amplifier	Heatsinks for 35W amplifier
TR_4 and TR_5	None	None	Thermal resistance 72 deg C/W; 2cm square 1mm bright aluminium
TR_6 and TR_7	Thermal resistance 21 deg C/W; 4cm square	Thermal resistance 12·5 deg C/W; 5·5cm square	Thermal resistance 7·25 deg C/W; 8cm square

Protection Circuits

Transistors and diodes

Circuit reference	Mullard type 25W amplifier	Mullard type 35W amplifier
TR_8	BC148	BC148
TR_9	BC158	BC157
D_1	BA145	BA148
D_2	BA145	BA148

Resistors

Circuit reference	Value 25W amplifier	Value 35W amplifier
R_{21}	8·2kΩ	10kΩ
R_{22}	8·2kΩ	10kΩ
R_{23}	4·7kΩ trimming potentiometer	4·7kΩ trimming potentiometer
R_{24}	4·7kΩ trimming potentiometer	4·7kΩ trimming potentiometer

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High-quality pre-amplifier

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UDC 621.375.021.029.45

A high-quality pre-amplifier, designed for use with the power amplifiers presented in the previous article, is described.

Fig. 1 shows a high-quality pre-amplifier using four plastic-encapsulated Lockfit transistors of the BC147 family. With a supply voltage of +30V, this circuit may be used with power amplifiers of about 400mV sensitivity. High a.c. and d.c. feedback ensures a very low distortion level (less than 0·15%) and minimises the effects of component parameter spreads. The output impedance is less than 1kΩ, and the –1dB frequency response, with tone controls in the flat position, is 10Hz to 50kHz. Bass and treble control potentiometers can be adjusted independently to give about ±19dB variation.

The circuit has five inputs—for magnetic and crystal pickups, radio tuner, tape recorder and magnetic microphone sources. The sensitivity in each position is given in the performance figures. An output signal for tape recording is also available.

Two pre-amplifiers may be used in stereo installations. The circuit diagram shows a left-hand channel, with points of connection to the right-hand channel.

CIRCUIT DESCRIPTION

The two input transistors TR_1 and TR_2, are low noise types BC149C and BC149B to reduce the noise output of the first two stages. The stages are directly coupled, with d.c. feedback to reduce the spread of transistor d.c. settings. A degree of a.c. feedback, varying for the different input sources, is also applied over these stages. In the magnetic pickup input circuit the first two stages are compensated in accordance with the R.I.A.A. norm by means of feedback network R_{16}, C_2, C_5. The remaining inputs have a frequency-independent feedback factor and flat frequency response. The overdrive margin of the first two stages is greater than 20dB referred to 350mV across the volume control potentiometer.

The voltage amplifications at 1kHz from the input to the top of the volume control potentiometer R_{19} are:

magnetic pickup	87·5 times
crystal pickup	1·16 times
radio tuner	2·33 times
tape recorder	1·16 times
magnetic microphone	100 times

Again at 1kHz, with nominal input signals, the voltage across the top of the volume control potentiometer is 350mV. This signal is fed through 1MΩ resistor R_{17} to a tape recorder output, so that a nominal recording signal of 350μV per 1kΩ of tape recorder input impedance is available.

The slider of the volume control potentiometer is connected to the base of transistor TR_3, type BC148B, which operates as an emitter follower. In this configuration the transistor output impedance is low, and the tone control circuit is fed from a voltage source and is independent of the position of the volume control slider.

The bass and treble controls are connected in the collector-to-base feedback loop of output transistor TR_4, type BC147B. This feedback system gives a low transistor output impedance and allows the tone controls to be varied independently over wide ranges. When the sliders of the tone control potentiometers R_{25} and R_{28} are in the flat position, the feedback is virtually independent of frequency and the voltage gain of the transistor is 1·35 times. The output signal from TR_4 is taken through resistors R_{36} and R_{35} and through balance control potentiometer R_{32}. The pre-amplifier output is taken from the junction of R_{35} and R_{36}.

With the slider of the balance control potentiometer in the centre position, the decrease in voltage amplification is about 0·6dB. For stereo operation the slider is taken to earth. For mono operation the slider is not earthed, and the inputs of both power amplifiers are connected to the pre-amplifier output.

The supply voltage is +30V. The first two stages have a decoupling network, R_{18} and C_3, the value of R_{18} being 4·7kΩ for mono and 2·7kΩ for stereo operation.

Typical d.c. settings (collector currents) for the transistors, at a nominal supply voltage of 30V, are:

TR_1 (BC149C)	75μA
TR_2 (BC149B)	730μA
TR_3 (BC148B)	3·5mA
TR_4 (BC147B)	2·4mA

The total current consumption is about 7mA.

PERFORMANCE

In the table the input sensitivity, input impedance, –1dB frequency response and signal to-noise ratio are given for each input. Sensitivity and input impedance are quoted for a frequency of 1kHz and for an output signal level of 440mV. The signal-to-noise ratios are given for an output signal level 20dB below 440mV.

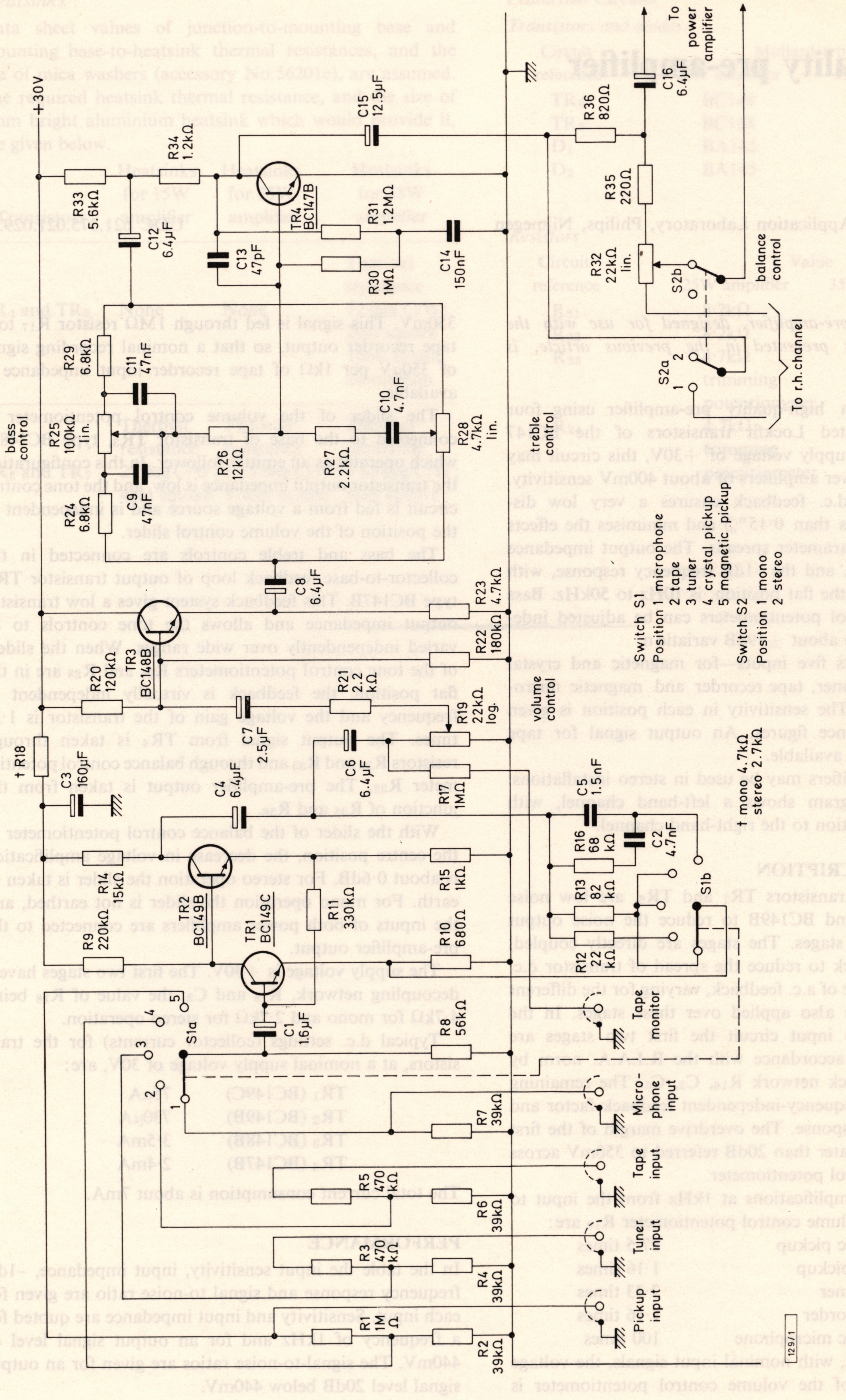
+30V
R33 5.6kΩ
R34 1.2kΩ
C15 12.5μF
R36 820Ω
C16 6.4μF
To power amplifier
C12 6.4μF
TR4 BC147B
R31 1.2MΩ
C13 47pF
R30 1MΩ
C14 150nF
R35 220Ω
R32 22kΩ lin.
S2b
balance control
S2a
To r.h. channel
R29 6.8kΩ
C11 47nF
C10 4.7nF
R28 4.7kΩ lin.
bass control
treble control
R25 100kΩ lin.
R26 12kΩ
R27 2.2kΩ
R24 6.8kΩ
C9 47nF
C8 6.4μF
R23 4.7kΩ
Switch S1 Position 1 microphone 2 tape 3 tuner 4 crystal pickup 5 magnetic pickup
Switch S2 Position 1 mono 2 stereo
TR3 BC148B
R20 120kΩ
R21 2.2kΩ
R22 180kΩ
C7 2.5μF
R19 22kΩ log.
volume control
† R18
C3 160μF
C4 6.4μF
C6 6.4μF
R17 1MΩ
C5 1.5nF
† mono 4.7kΩ stereo 2.7kΩ
R16 68kΩ
C2 4.7nF
S1b
R15 1kΩ
R14 15kΩ
TR2 BC149B
TR1 BC149C
R11 330kΩ
R10 680Ω
R13 82kΩ
R12 22kΩ
R9 220kΩ
C1 1.6μF
S1a
R8 56kΩ
Tape monitor
R7 39kΩ
Microphone input
Tape input
R5 470kΩ
R6 39kΩ
Tuner input
R3 470kΩ
R4 39kΩ
R1 1M
Pickup input
R2 39kΩ
129/1

Fig. 1—Pre-amplifier circuit

Input	Sensitivity (mV)	Input impedance (kΩ)	Frequency response	Signal-to-noise ratio (dB)
magnetic pickup	4	47	see Fig. 2	>90
crystal pickup	300	1000	10Hz to 35kHz	>80
radio tuner	150	500	10Hz to 35kHz	>80
tape recorder	300	500	10Hz to 45kHz	>85
magnetic microphone	3·5	22	10Hz to 65kHz	>80

The overdrive margin of the circuit is greater than 20dB with respect to the nominal output voltage of 440mV. Thus the undistorted output voltage is 4·4V.

The output impedance of the pre-amplifier is less than 1kΩ.

The total harmonic distortion is less than 0·15% between 20Hz and 20kHz for all inputs and for the nominal output voltage.

The available tape recorder signal is 350μV per 1kΩ load impedance.

The tone control characteristics of the circuit are shown in Fig. 3.

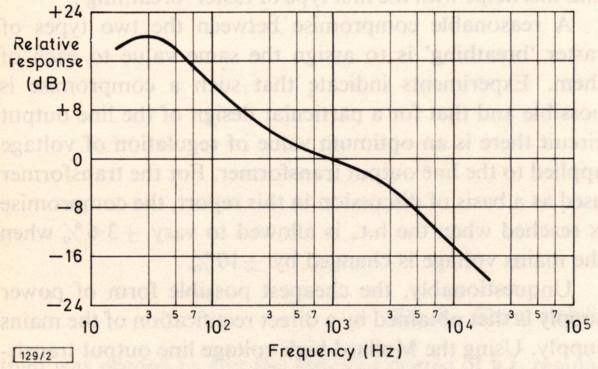

Fig. 2—Frequency characteristic (magnetic pickup input)

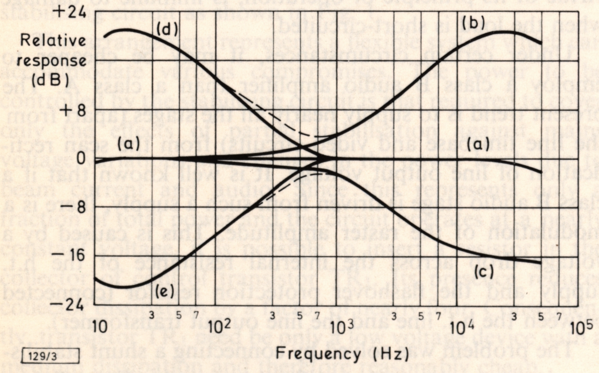

Fig. 3—Tone control characteristics
(a) bass and treble controls flat;
(b) maximum treble, bass flat;
(c) minimum treble, bass flat;
(d) maximum bass, treble flat;
(e) minimum bass, treble flat.
The dotted line joining (b) and (d) indicates maximum bass, maximum treble, and that joining (c) and (e) indicates minimum bass, minimum treble.

COMPONENTS

Transistors

Circuit reference	Mullard type	Circuit reference	Mullard type
TR_1	BC149C	TR_3	BC148B
TR_2	BC149B	TR_4	BC147B

Resistors ($\frac{1}{8}$W, ±10% tolerance)

Resistors R_9, R_{11} and R_{30} should be low noise components.

Circuit reference	Value	Circuit reference	Value
R_1	1MΩ	R_{19}	22kΩ logarithmic potentiometer
R_2	39kΩ	R_{20}	120kΩ
R_3	470kΩ	R_{21}	2·2kΩ
R_4	39kΩ	R_{22}	180kΩ
R_5	470kΩ	R_{23}	4·7kΩ
R_6	39kΩ	R_{24}	6·8kΩ
R_7	39kΩ	R_{25}	100kΩ linear potentiometer
R_8	56kΩ	R_{26}	12kΩ
R_9	220kΩ	R_{27}	2·2kΩ
R_{10}	680Ω	R_{28}	4·7kΩ linear potentiometer
R_{11}	330kΩ	R_{29}	6·8kΩ
R_{12}	22kΩ	R_{30}	1MΩ
R_{13}	82kΩ	R_{31}	1·2MΩ
R_{14}	15kΩ	R_{32}	22kΩ linear potentiometer
R_{15}	1kΩ	R_{33}	5·6kΩ
R_{16}	68kΩ	R_{34}	1·2kΩ
R_{17}	1MΩ	R_{35}	220Ω
R_{18}	4·7kΩ mono 2·7kΩ stereo	R_{36}	820Ω

Capacitors

Circuit reference	Value	Description		Mullard type
C_1	1·6μF	electrolytic	25V	C426AS/F1·6
C_2	4·7nF	polyester		C296AC/A4K7
C_3	160μF	electrolytic	25V	C437AR/F160
C_4	6·4μF	electrolytic	25V	C426AR/F6·4
C_5	1·5nF	polyester		C296AC/A1K5
C_6	6·4μF	electrolytic	25V	C426AR/F6·4
C_7	2·5μF	electrolytic	16V	C426AR/H2·5
C_8	6·4μF	electrolytic	25V	C426AR/F6·4
C_9	47nF	metallised film		C280AE/P47K
C_{10}	4·7nF	polyester		C296AC/A4K7
C_{11}	47nF	metallised film		C280AE/P47K
C_{12}	6·4μF	electrolytic	25V	C426AR/F6·4
C_{13}	47pF	ceramic plate		C333CC/C47E
C_{14}	150nF	metallised film		C280AE/P150K
C_{15}	12·5μF	electrolytic	25V	C426AR/F12·5
C_{16}	6·4μF	electrolytic	25V	C426AR/F6·4

A simple stabilised power supply for transistorised monochrome receivers

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UDC 621.397.13: 621.311.62—503

In this article it is established that a satisfactory operation of a fully transistorised monochrome receiver can be obtained using a partially stabilised h.t. supply. The optimum conditions can be reached by adjustment of the characteristic of the supply to that of the line output transformer. A shunt stabilising circuit connected to the scan rectification supply via a series resistor can provide the required degree of stabilisation and at the same time serve as a source of power for class B audio amplifier. Partial stabilisation and low voltage operation (with a resistor in collector of the shunt stabilising transistor) contribute to the low cost of the circuit.

Although the proposed circuit has a built-in immunity to various fault conditions, in common with other feedback circuits (including hybrid receivers) it produces a large increase in voltage when an open-circuit occurs.

INTRODUCTION

This article examines the requirements of a power supply for a monochrome receiver employing a high voltage transistor in the line timebase, and describes an economical solution to these requirements.

Although the development work connected with this project was done using a television receiver operated at 20kV and a class A field output timebase, the ideas evolved can be applied to any other execution. For instance, the calculations shown in Appendix 2 indicate that a worthwhile reduction in power consumption (and perhaps some saving in cost) may be obtained with a receiver designed for 18kV operation with class B field output stage.

UNDERLYING PRINCIPLES

It has been established experimentally that the e.h.t. regulation associated with a transistorised line timebase is not as good as that of corresponding e.h.t. supply in a line timebase using valves. This results in a worse performance in respect of the so-called raster 'breathing', that is, a change in raster size with a change of current taken from the e.h.t. supply by the picture tube. A simple way of compensating for this deficiency is to introduce some resistance into the h.t. supply. Although this improves raster 'breathing' and the viewer is less aware of changes, the e.h.t. regulation is worsened.

Thus, a possible design for a television receiver would be a stabilised h.t. supply with some series resistance. However, once a certain amount of raster 'breathing' as a function of beam current is accepted, it is logical to extend this principle to raster 'breathing' as a function of mains voltage variations. This implies that the power supply need not be so well stabilised and therefore it may cost less. As it happens, a concession on the amount of regulation leads to an increase in the internal resistance of the power supply and this helps with the first type of raster 'breathing'.

A reasonable compromise between the two types of raster 'breathing' is to assign the same value to each of them. Experiments indicate that such a compromise is possible and that for a particular design of the line output circuit there is an optimum value of regulation of voltage applied to the line output transformer. For the transformer used as a basis of discussion in this report, the compromise is reached when the h.t. is allowed to vary ±3·6% when the mains voltage is changed by ±10%.

Unquestionably, the cheapest possible form of power supply is that obtained by a direct rectification of the mains supply. Using the Mullard high voltage line output transistor BU105, a reasonable value of h.t. is between 130 and 150V. Unfortunately, this implies a considerable amount of power wasted. However, this provides a favourable condition for the application of shunt stabilisation which, by virtue of its principle of operation, is immune to damage when the load is short-circuited.

Under certain circumstances, it may be cheaper to employ a class B audio amplifier than a class A. The present trend is to supply nearly all the stages (apart from the line timebase and video circuits) from the scan rectification of line output voltage. It is well known that if a class B audio stage is driven from such a supply, there is a modulation of the raster amplitude. This is caused by a voltage drop across the internal resistance of the h.t. supply and the flashover protection resistor (connected between the h.t. line and the line output transformer).

The problem was solved by connecting a shunt stabilising circuit to the scan rectification supply. For efficient operation a good coupling is required between the scan rectification winding and the main winding. Since such a coupling is needed in any case to ensure good regulation and a minimum current at the end of scan in the scan rectification diode, this requirement is already incorporated when a scan rectification supply is used. Furthermore, the shunt stabilising circuit was designed for a lower impedance than was needed to give the required degree of h.t. regulation and a series resistor (R_3 in Fig. 1) was provided to make up the difference. While the power to low voltage stages of the receiver was taken directly from the scan rectification supply, the class B audio amplifier was fed from the shunt stabilising circuit as shown in Fig. 1.

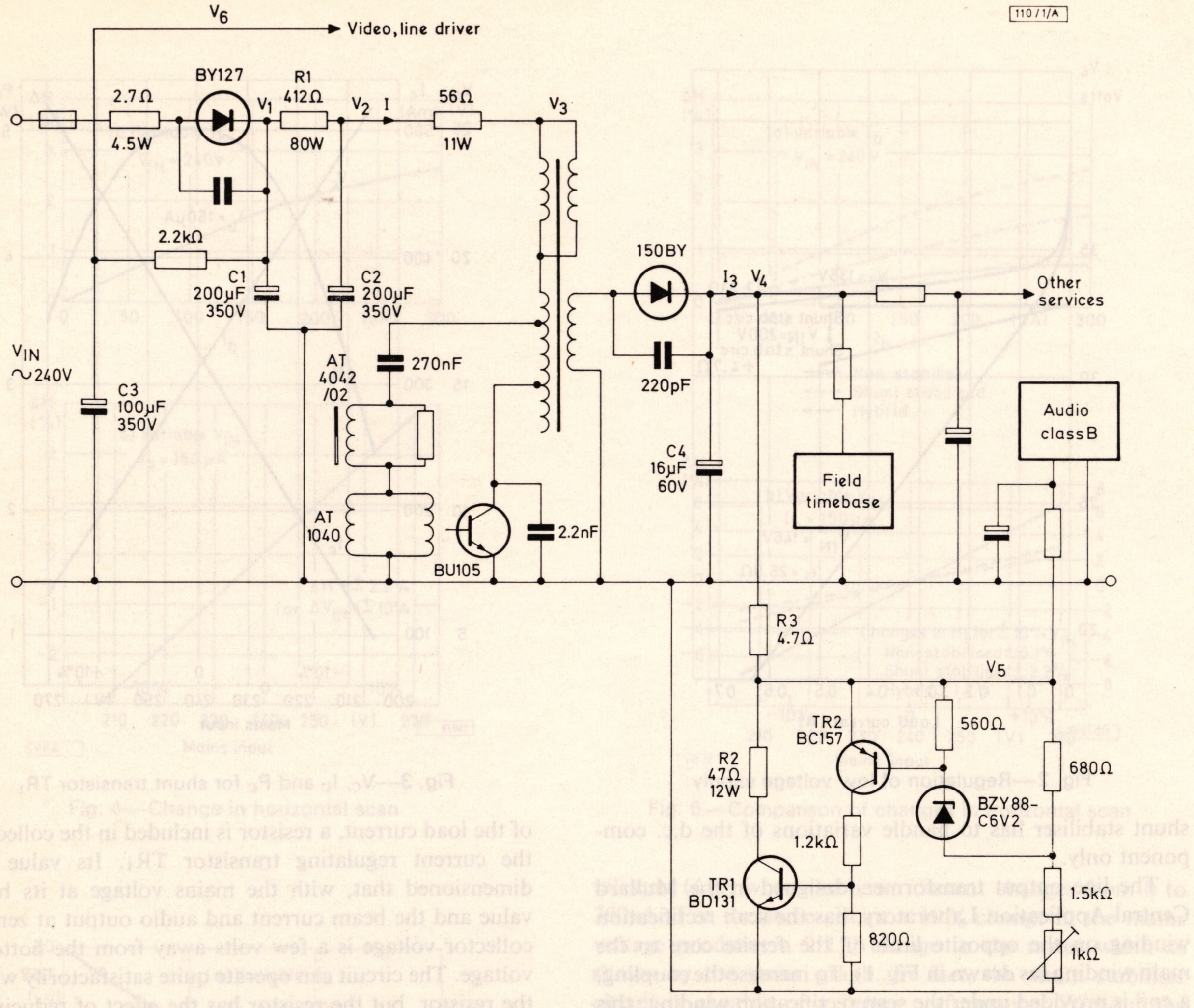

Fig. 1—Shunt stabilised scan rectification supply

This arrangement represents a flexible system which can accommodate various compromises. The power to be controlled by the stabilising circuit is that required to cover only the effects of partial stabilisation against mains voltage variations and changes in the power levels due to beam current and audio. Since this represents only a fraction of total power and the circuit operates at a nearly constant voltage, it is possible to insert a resistor in the collector of control transistor TR_1. Its presence reduces collector dissipation by a factor of nearly four. Consequently, transistor TR_1 need be only a low voltage device with a medium dissipation and therefore reasonably cheap.

CIRCUIT DESCRIPTION

H.T. supply

As shown in Fig. 1, a single-phase rectifier provides h.t. supply to the receiver. Resistor R_1 serves as a voltage dropper and in conjunction with C_2 forms R–C smoothing for the supply. It is followed by a 56Ω resistor (without any decoupling) whose sole purpose is to provide flashover protection for the line timebase.

The h.t. for the video stage and the line output transistor drive is obtained directly from the reservoir capacitor C_1. The purpose of this arrangement is to remove a large increase in the h.t. value in the event of the line timebase becoming open-circuited.

Scan rectification supply

All the remaining stages of the receiver are supplied from one low voltage supply obtained by scan rectification. Its value is determined primarily by the requirement of the field output stage. Only one scan rectification diode need be used. The new Mullard diode 150BY has adequate current carrying capacity for this application. The value of the reservoir capacitor is kept deliberately low in order to provide a short time constant. It was noted that with larger values there was an overshoot in the raster 'breathing' on a sudden change in the beam current. In consequence the branches of the low voltage line feeding various stages carry their own decouplings.

As mentioned earlier, the audio output stage, operated in class B, is connected directly to the shunt stabiliser. It carries its own decouplings for a.c. currents so that the

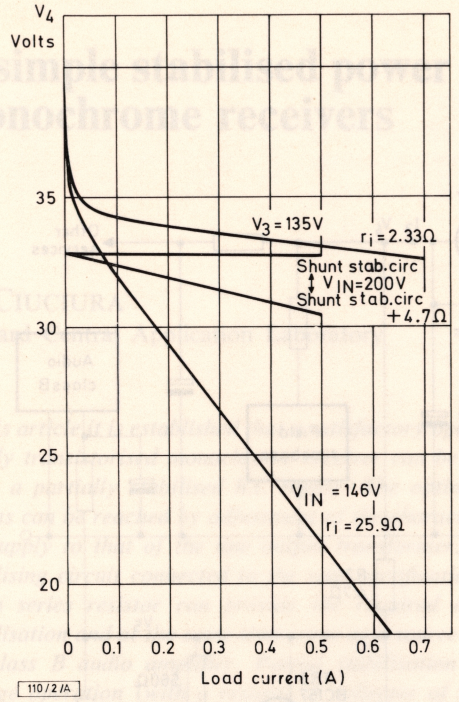

Fig. 2—Regulation of low voltage supply

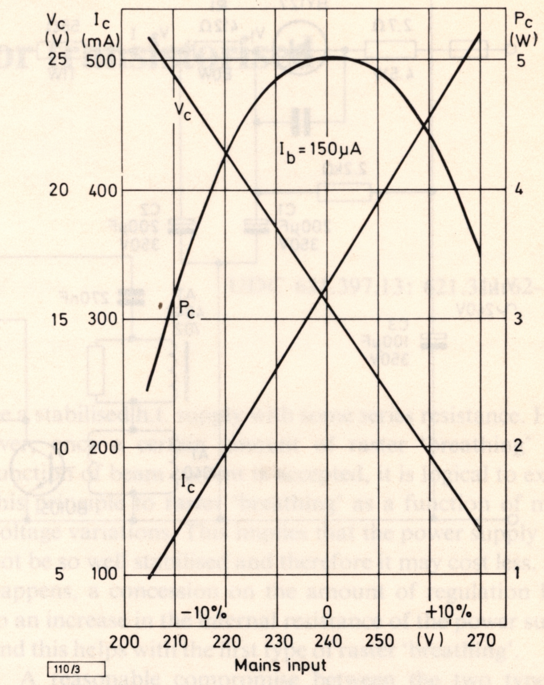

Fig. 3—V_C, I_C and P_C for shunt transistor TR$_1$

shunt stabiliser has to handle variations of the d.c. component only.

The line output transformer, designed in the Mullard Central Application Laboratory, has the scan rectification winding on the opposite limb of the ferrite core to the main winding (as drawn in Fig. 1). To increase the coupling, a coil is provided under the scan rectification winding; this coil is connected across an appropriate number of turns on the main winding.

In Fig. 2 the curve for constant $V_{IN} = 146$V shows the regulation of the scan rectification supply on its own. For currents higher than 0·1A the graph is nearly a straight line representing a resistance of about 30 Ω. This resistance consists of the reflected internal resistance of the h.t. line plus the intrinsic resistance of the line output transformer and scan rectification diode. The sum of the latter two impedances is shown in Fig. 2 as a graph taken at a constant voltage input to the line output transformer ($V_3 = 135$V).

Both graphs show a sharp increase in voltage for load currents of, say, less than 30mA. This is due to a spike on the scan rectification waveform following the flyback pulse.

Shunt stabiliser

The circuit diagram of the shunt stabiliser is shown in the lower portion of Fig. 1. It is a fairly conventional arrangement and one of many possible solutions. Transistor TR$_1$ forms a shunt across the scan rectification supply. Its collector current is controlled by transistor TR$_2$. With increase in the supply voltage, more current is passed by transistor TR$_2$; this increases the base current of transistor TR$_1$, thus increasing the collector current of this transistor. This leads to a greater loading of the h.t. line and thus to stabilisation of the low voltage supply.

Since the power supply has to deal only with a portion of the load current, a resistor is included in the collector of the current regulating transistor TR$_1$. Its value is so dimensioned that, with the mains voltage at its highest value and the beam current and audio output at zero, the collector voltage is a few volts away from the bottoming voltage. The circuit can operate quite satisfactorily without the resistor, but the resistor has the effect of reducing the collector dissipation by a factor of nearly four. This obviously has an effect on the size of the transistor and the associated heat sink.

The behaviour of the transistor TR$_1$ can be easily understood with the aid of graphs in Fig. 3. With the increase in the mains voltage, there is an increase in collector current and a corresponding drop in the collector voltage. At the same time, the dissipation of the transistor initially increases, reaches its maximum at about nominal mains voltage, and then decreases.

PERFORMANCE

The aim of this project was to provide the same performance of the line timebase (and the television receiver) using transistors as is normally obtainable with hybrid receivers. In going from one technique to the other, not only the stabilising circuit is changed, but also the performance of the line timebase. It is well known that for the same configuration of the line output transformer, the transistorised execution produces e.h.t. of a higher internal impedance than the hybrid execution. However, in so far as a complete receiver is concerned, the only factors that matter are changes in raster size with beam current (breathing) and with the mains voltage. The performance of the circuit shown in Fig. 1 is given in the form of graphs in Figs. 4 and 5. The top graph shows the change in the raster size with beam current. There is a 2·6% increase in the horizontal scan (ΔH) when the beam current changes from 0 to 300μA. For the same circuit a 10% change in the mains voltage produces a 2·2% change in the raster width as shown by the lower graph. In fact, the shunt stabiliser circuit is capable of a better performance. Without the resistor R$_3$ in series with the shunt stabiliser, there is only 1·5% change in the raster width for 10% change in the mains voltage, but the raster breathing increases to 3·5% (0 to 300μA). Thus, the purpose of the series resistor R$_3$ is to provide a reasonable compromise between the raster breathing and its change with the mains voltage.

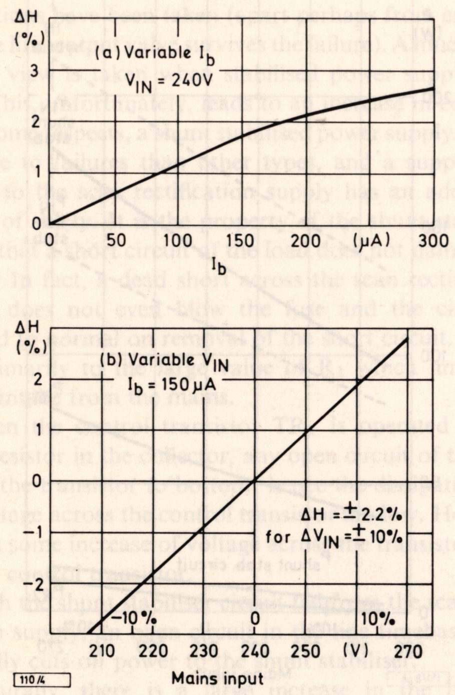

Fig. 4—Change in horizontal scan

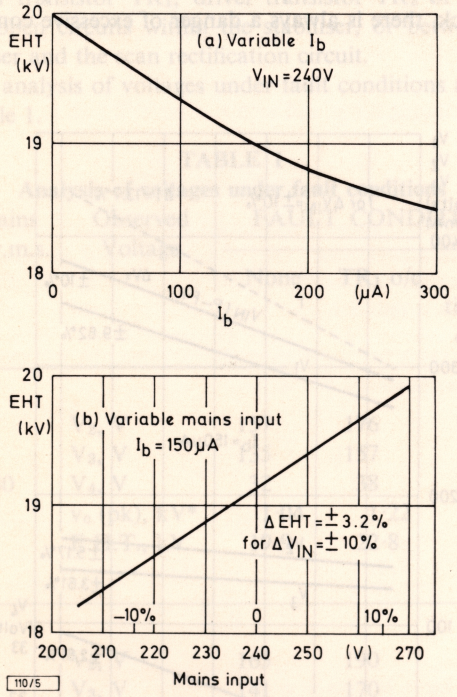

Fig. 5—Characteristics of e.h.t. supply

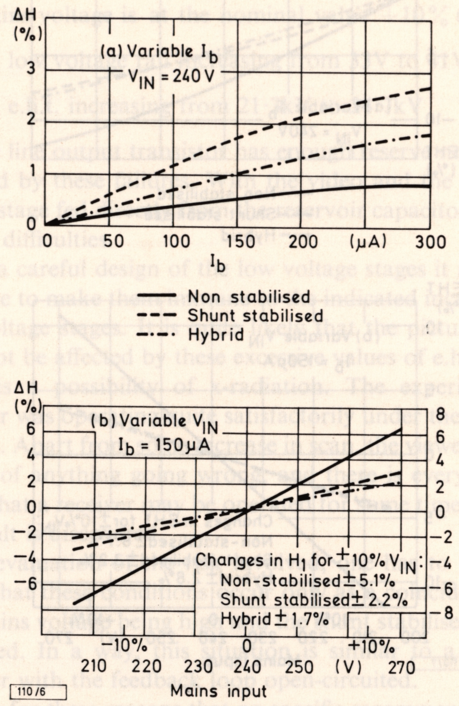

Fig. 6—Comparison of changes in horizontal scan

It is of interest to compare the performance of the proposed circuit with a hybrid receiver. Initial investigation showed quite a large spread in performance of hybrid receivers of various brands. On the assumption that the hybrid technique is capable of good performance, one of the best available receivers was selected for the purpose of comparison. The results are shown in Fig. 6. The lower portion in this figure compares behaviour of three receivers as a function of mains voltage. There is very little difference (0·5%) between the hybrid receiver and the shunt stabilised receiver; a non-stabilised transistorised receiver shows 5% change in raster width for 10% change in mains voltage.

In respect of raster 'breathing' with beam current, the shunt stabilised receiver has a somewhat worse performance than the hybrid receiver (2·6% as compared with 1·8% for 0 to 300μA change in beam current). The best performance is given by the non-stabilised receiver (0·8%).

The above comparison can be carried out in terms of e.h.t. changes and the relevant results are shown in Fig. 7. There the difference between the hybrid and the shunt stabilised receivers is smaller than in Fig. 6.

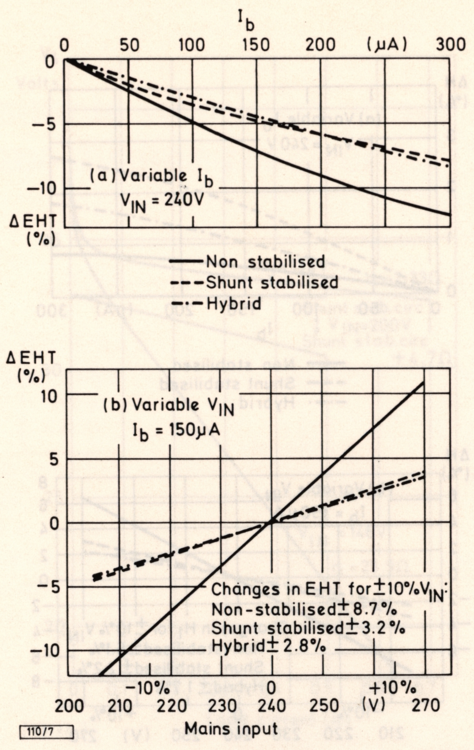

Fig. 7—Comparison of change in e.h.t. supply

POWER CONSUMPTION

Since the voltage required by the line timebase is well below the value obtained by rectification of mains voltage, a considerable proportion of power is dissipated in the voltage dropping resistor. Nevertheless, the receiver requires about 35W less power than a hybrid receiver, as can be seen in Fig. 8. The values quoted were measured at the mains input to the receiver using a dynamometer type wattmeter.

The largest consumer of power is in fact the voltage dropping resistor. With normal mains voltage it dissipates 57W, and with mains voltage increased to +10%, 73W. Provisions must be made for the disposal of this power. Fortunately, this is not a critical component in respect of location, and a suitable position can usually be found inside the cabinet where the heat generated can be easily conducted away.

Another resistor with a significant amount of dissipation is R_2 in series with the shunt regulator transistor. With mains voltage at +10% above nominal, its dissipation is 16W. Its positioning may be more critical than that of R_1, since it forms a part of the feedback circuit.

Behaviour of various voltages as a function of mains voltage variation is shown in Fig. 9 (for changes in e.h.t. see Figs. 5 and 7).

It is necessary to stress that the above data apply to a receiver operated at 20kV and using a class A field output stage. As mentioned earlier, a worthwhile reduction in power consumption is obtained with 18kV design and class B field timebase (see Appendix 2). In such a receiver the input power is 110W and the dissipation in R_1 is 97W at nominal mains voltage.

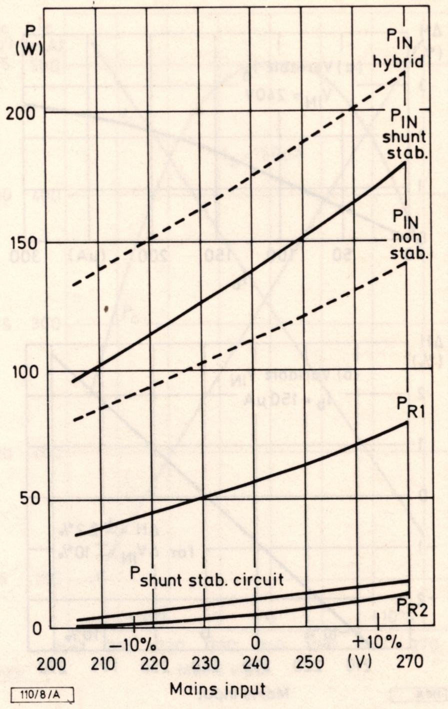

Fig. 8—Mains input power, and dissipation of resistors

PROTECTION AGAINST FAILURES

In any system where power is controlled by means of feedback, there is always a danger of excessive conditions when, for any reason, there is a breakdown in the feedback loop. This problem also exists in hybrid receivers, but few precautions have been taken (apart perhaps from ensuring that the line output valve survives the failure). A much more serious view is taken when stabilised power supplies are used. This, unfortunately, leads to an increase in cost.

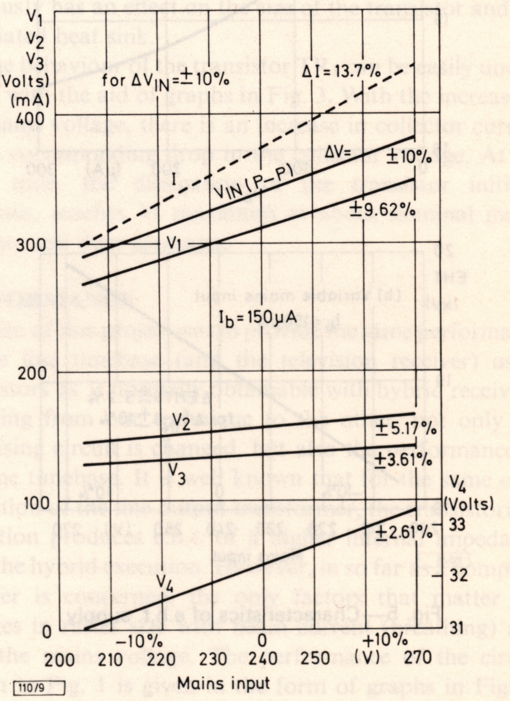

Fig. 9—Variation of voltages with mains supply

In some respects, a shunt stabilised power supply is more immune to failures than other types, and a supply connected to the scan rectification supply has an additional degree of safety. It is the property of the shunt stabilised circuit that a short circuit of the load does not damage the supply. In fact, a dead short across the scan rectification supply does not even blow the fuse and the circuit is restored to normal on removal of the short circuit. This is due primarily to the large value of R_1 which limits the power intake from the mains.

When the control transistor TR_1 is operated with a series resistor in the collector, any open circuit of the load causes the transistor to bottom; hence the dissipation and the voltage across the control transistor are low. However, there is some increase of voltage across the transistor feeding the control transistor.

With the shunt stabiliser circuit fed from the scan rectification supply, an open circuit in the line timebase automatically cuts off power to the shunt stabiliser.

Naturally, there is a large increase in the h.t. line potential but, with other stages supplied directly from the reservoir capacitor, there is no danger.

The only remaining difficulty is when the shunt stabiliser is put out of action so that it does not draw any current from the scan rectification supply. This can be caused by control transistor TR_1, driver transistor TR_2 or several other open circuits within the stabiliser, or between the stabiliser and the scan rectification circuit.

An analysis of voltages under fault conditions appears in Table 1.

TABLE 1

Analysis of voltages under fault conditions

Mains V, r.m.s.	Observed Voltage	FAULT CONDITIONS None	TR_1 o/c	Line timebase o/c
240	V_2, V	157	176	327
	V_3, V	135	157	327
	V_4, V	32	38	—
	v_c (pk), kV*	1·04	1·22	—
	E.H.T., kV	19·9	22·8	—
264	V_2, V	165	190	353
	V_3, V	141	170	353
	V_4, V	33·1	41·2	—
	V_c (pk), kV*	1·08	1·32	—
	E.H.T., kV	21·3	24·7	—

*peak voltage on the line output transistor, BU105
E.H.T. measured at I_b = 0

Table 1 indicates that there are two dangers when the shunt stabilised power supply becomes open-circuited and the mains voltage is at the nominal value +10% (264V):

a) low voltage rail increasing from 33V to 41V.

b) e.h.t. increasing from 21·3kV to 24·7kV

The line output transistor has enough reserve not to be affected by these failures. With the video and the BU105 driver stage fed directly from the reservoir capacitor, there are no difficulties.

By a careful design of the low voltage stages it may be possible to make them immune to the indicated increase in low voltage stages. It is quite likely that the picture tube may not be affected by these excessive values of e.h.t., but there is a possibility of x-radiation. The experimental receiver was operating quite satisfactorily under these conditions. Apart from some increase in scan, the viewer is not aware of anything going wrong, and there is every likelihood that a receiver may be operated for some time before the fault is discovered.

In evaluation of the risk involved, one has to bear in mind that these conditions occur only as a coincidence of the mains voltage being high and the shunt stabiliser open-circuited. In a way, this situation is similar to a hybrid receiver with the feedback loop open-circuited.

It is for these reasons that no specific recommendations are offered in this article and the decision is left to the set designer. Suitable methods of protection will be offered in a later article. The risk of x-radiation from the picture tube may be eliminated by reduction of e.h.t. to 80% of the Design Centre maximum value (18kV operation).

DESIGN NOTES

The operation of the shunt stabiliser circuit is intimately connected with the performance of the h.t. supply; in fact the two circuits are parts of the same unit. Because of the inevitable interaction, designers confronted for the first time with this type of power supply may find it somewhat confusing. It may be helpful to suggest the following points:

a) The voltage applied to the line output transformer (V_3) is decided by the design of the transformer. The object of the power supply is to provide this voltage. Adjustment can be made by altering the value of R_1, taking care that there is a required amount of current through the shunt stabiliser.

b) The value of scan rectification voltage at the correct load can be adjusted only by changing the number of turns on the line output transformer.

c) In selection of current for the shunt stabiliser, care must be taken about the effects of mains voltage variations, component tolerances and operational changes due to beam current, audio output, a.g.c. and so on.

A theoretical analysis of some aspects of the shunt stabiliser operation is given in Appendix 1. This is followed by design considerations in Appendix 2. The values provided by these calculations, while not accurate, should form a good basis for experimental work.

APPENDIX 1

Effects of voltage and load changes on shunt stabilising circuit

A shunt stabiliser circuit can be represented by an equivalent network as shown in Fig. 10. Here an h.t. line at potential V supplies power to the load R_L via series resistor R. The stabilising action is produced by the current generator I_S. This generator is controlled in such a way that at the minimum value of h.t. for which the system is designed its current is zero. With

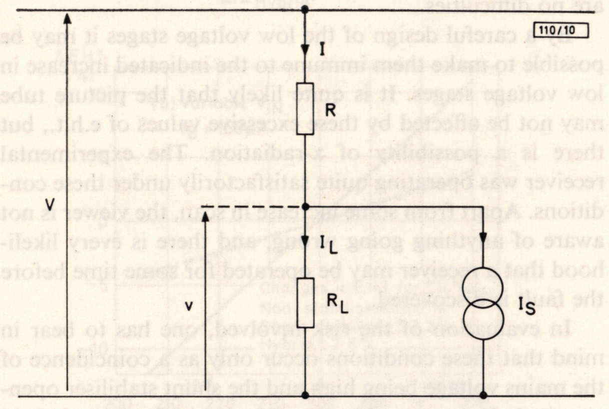

Fig. 10—Equivalent network of shunt stabiliser

the increase in the h.t. line voltage, the generator passes a controlled amount of current so that the rate at which the voltage increases across the load is smaller than across the h.t. line.

Let the allowable variation in the h.t. voltage be pV, and the corresponding variation in the load voltage xv.

Let $V = kv$ where $k > 1$

The value of the load R_L is set by the circuit requirements under normal operating conditions. However, the value of R is selected so that when h.t. reaches its minimum design value, the current in the shunt stabiliser is zero. The voltage distribution then is shown in Fig. 11.

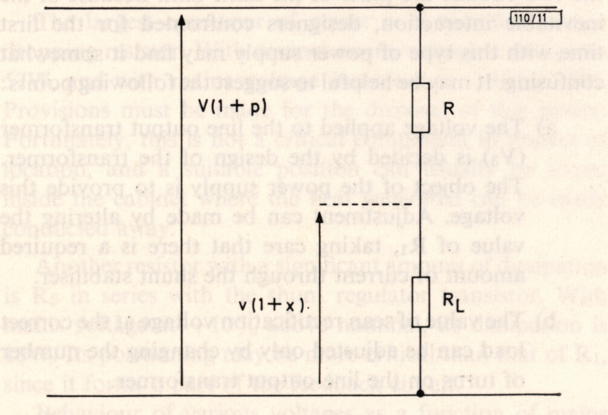

Fig. 11—Voltage distribution in network at minimum h.t.

Under these conditions:

$$\frac{R}{R_L} = \frac{V(1-p)-v(1-x)}{v(1-x)} = \frac{k(1-p)-(1-x)}{1-x} = r \quad \ldots(1)$$

Under normal operating conditions the shunt stabiliser passes current; its value is equal to $I-I_L$. With reference to Fig. 10,

$$I = \frac{V-v}{R} = \frac{v}{R_L}\cdot\frac{k-1}{r}.$$

$$I_L = \frac{v}{R_L}.$$

Hence,

$$I_S = \frac{v}{R_L}\left(\frac{k-1}{r}-1\right) = z.I_L \quad \ldots(2)$$

Therefore, the power taken from the h.t. line is given by

$$P_{IN} = VI = \frac{v^2}{R_L}.k.\frac{k-1}{r}.$$

Power in the load is given by

$$P_L = v.I_L = \frac{v^2}{R_L},$$

and the power in the shunt stabiliser is given by

$$P_S = vI_s = \frac{v^2}{R_L}\cdot\left(\frac{k-1}{r}-1\right) = z.P_L$$

where

$$z = \frac{k-1}{r}-1.$$

Substituting for r gives

$$z = \frac{(k-1)(1-x)}{k(1-p)-(1-x)}-1 = \frac{k(p-x)}{k(1-p)-(1-x)}. \quad \ldots(3)$$

In practice the expected excursions of unstabilised h.t. line are within ±10%, hence $p = 0{\cdot}1$.

$$\text{Let } \frac{x}{p} = q = \frac{\%\text{ change of voltage across load}}{\%\text{ change of h.t. voltage}}$$

where $1 \geqslant q \geqslant 0$.

Substituting for p, and with $x = pq = 0{\cdot}1 \times q$,

$$z = \frac{(0{\cdot}1)k(1-q)}{(0{\cdot}9)k+(0{\cdot}1)q-1} \quad \ldots(4)$$

In Fig. 12 is shown a family of graphs of z as a function of q with k as a parameter. They indicate a rapid increase in the power which the stabilising circuit has to handle when the value of k is reduced.

With the h.t. at its highest design value, the voltages in the circuit are as shown in Fig. 13.

$$I_2 = \frac{vk(1+p)-v(1+x)}{rR_L}$$

$$= \frac{v}{R_L}.\frac{k(1+p)-(1+x)}{r}$$

$$I_{L2} = \frac{v}{R_L}(1+x)$$

$$I_{S2} = I_2-I_{L2} = \frac{v}{R_L}\left[\frac{k(1+p)-(1+x)}{r}-(x+1)\right] \quad \ldots(5)$$

Since with a nominal value of h.t. the shunt stabilising circuit should have enough reserve of current to operate with h.t. at the minimum value, the same amount of reserve of current should be between h.t. line at the maximum value and the nominal conditions.

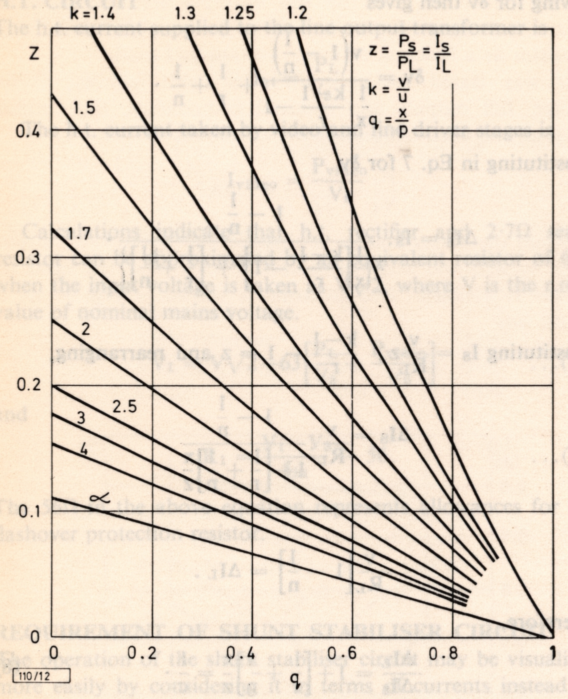

Fig. 12—Power ratio

In other words:

$$I_{S2} = 2I_S.$$

Comparing Eq. 2 and Eq. 5, this is not obvious. However, the equality can be proved by the following argument:

$$I_{S2} = \frac{v}{R_L}\left[\frac{k(1+p)-x-1}{r}-x-1\right]$$

$$= \frac{v}{R_L}\left[\frac{k+kp-x-1}{r}-x-1\right]$$

$$= \frac{v}{R_L}\left[\frac{k-1}{r}-1+\frac{kp-x}{r}-x\right]$$

$$= I_S+\frac{v}{R_L}\left(\frac{kp-x}{r}-x\right).$$

From Eq. 1,

$$\frac{k(1-p)-(1-x)}{1-x} = r.$$

Rearranging gives

$$x = 1-\frac{k(1-p)}{r+1}.$$

Substituting for x gives

$$I_{S2} = I_S+\frac{v}{R_L}\cdot\frac{(kp-1)(r+1)+k(1-p)}{r(r+1)}+\frac{rk(1-p)}{r(r+1)}-1$$

$$= I_S+\frac{v}{R_L}\cdot\frac{k(r+1)-(r+1)}{r(r+1)}-1$$

$$= I_S+\frac{v}{R_L}\cdot\frac{k-1}{r}-1$$

$$I_{S2} = I_S+I_S = 2I_S. \quad \ldots(6)$$

In practice the shunt stabilising circuit will be controlled by a voltage proportional to the voltage across the load. The

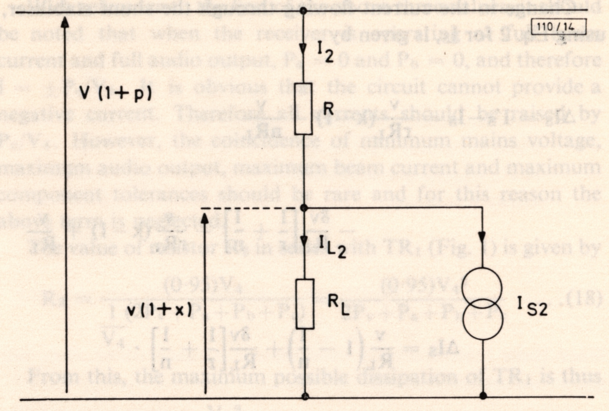

Fig. 13—Voltage distribution in network at maximum h.t.

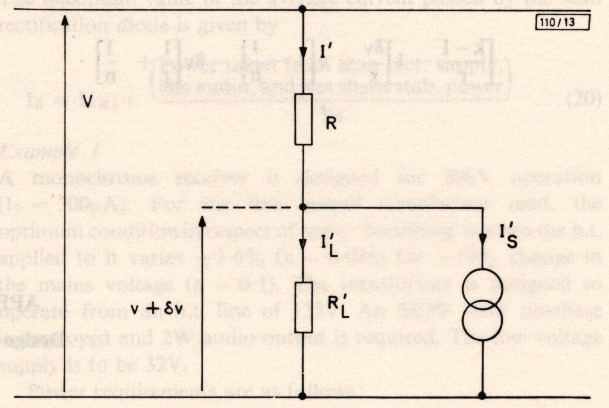

Fig. 14—Voltage distribution with reduced load

above equation indicates that for a change xv across the load, the current in the shunt stabiliser changes by I_S. Therefore for a change δv

$$\Delta I_S = I_S\frac{\delta v}{xv}. \quad \ldots(7)$$

EFFECT OF CHANGE OF LOAD

Let us assume that, with nominal mains voltage, the load has been reduced, so that the resistance representing the load has been increased to nR_L with $n > 1$. As a result of this, the voltage across the load is increased by δv. The new situation is shown in Fig. 14, and the corresponding current is as follows:

$$I' = \frac{V-v+\delta v}{R} = \frac{v(k-1)}{rR_L}-\frac{\delta v}{rR_L}$$

$$I'_L = \frac{v+\delta v}{R'_L} = \frac{v}{nR_L}+\frac{\delta v}{nR_L}$$

$$I'_S = I'-I'_L = \frac{v}{rR_L}.(k-1)-\frac{\delta v}{rR_L}-\frac{v}{nR_L}-\frac{\delta v}{nR_L}$$

Change in the current flowing through the shunt stabiliser, using Eq. 2 for I_S, is given by

$$\Delta I_S = I'_S - I_S = \frac{v}{rR_L}(k-1) - \frac{v}{nR_L}$$

$$-\frac{\delta v}{R_L}\left[\frac{1}{r}+\frac{1}{n}\right] - \frac{v}{rR_L}(k-1) + \frac{v}{R_L}$$

$$\Delta I_S = \frac{v}{R_L}\left(1-\frac{1}{n}\right) + \frac{\delta v}{R_L}\left[\frac{1}{r}+\frac{1}{n}\right].$$

Eq. 7 gives ΔI_S as

$$\Delta I_S = I_S\frac{\delta v}{xv} = \frac{v}{R_L}\left[\frac{k-1}{r}-1\right]\frac{\delta v}{xv}.$$

Equating and dividing both sides by R_L gives,

$$\left[\frac{k-1}{r}-1\right]\frac{\delta v}{x} = v\left[1-\frac{1}{n}\right] - \delta v\left[\frac{1}{r}+\frac{1}{n}\right].$$

Solving for δv then gives

$$\delta v = \frac{v\left(1-\frac{1}{n}\right)}{\frac{1}{x}\cdot\frac{k-1}{r}-1} + \frac{1}{r} + \frac{1}{n}.$$

Substituting in Eq. 7 for δv

$$\Delta I_S = I_S.\frac{1-\frac{1}{n}}{x\left\{\frac{1}{x}\left[\frac{k-1}{r}-1\right]+\left[\frac{1}{r}+\frac{1}{n}\right]\right\}}.$$

Substituting $I_S = \frac{v}{R_L}z$, $\frac{k-1}{r} - 1 = z$ and rearranging,

$$\Delta I_S = \frac{v}{R_L}.\frac{1-\frac{1}{n}}{1+\left[\frac{1}{r}+\frac{1}{n}\right]\frac{x}{z}}$$

but

$$\frac{v}{R_L}\left[1-\frac{1}{n}\right] = \Delta I_L.$$

Therefore,

$$\frac{\Delta I_L}{\Delta I_S} = 1+\left[\frac{1}{r}+\frac{1}{n}\right]\frac{x}{z} = \frac{1}{c}. \qquad \ldots(8)$$

APPENDIX 2

Design considerations

ASSESSMENT OF H.T. POTENTIAL

The BU105 line output transistor has a peak collector voltage rating of 1·5kV. Allowing 20% of v_c for various component spreads, loss of sync, h.t. variations and so on, the design centre value is 1·25kV. If a further allowance of 25% is made to safeguard the transistor against any possibility of an open circuit within the shunt stabilised power supply, the workable value of collector voltage is 1kV.

From standard formulae, the value of h.t. voltage V_3 is:

$$V_3 = \frac{v_c(pk)}{(1-a)(1+Fp)} \qquad \ldots(9)$$

where $Fp = \frac{\pi}{2}.\frac{1-p}{p} + \frac{2}{\pi}$, and $p = \frac{\tau f}{\tau s}$ – flyback ratio

and where a is a reduction in peak voltage due to harmonic tuning.

Taking $p = 0{\cdot}18$, $F_p = 7{\cdot}8$ and $a = 0{\cdot}15$, then

$$V_3 = 134V.$$

POWER REQUIREMENTS

It is assumed that the power to the video output stage and the line driver stage (10·5W) will be taken directly from the reservoir capacitor and for this reason will be excluded from calculations.

The process of calculation necessitates establishing the amount of power P to be handled by the line output transformer at full output and nominal mains voltage. From the value of P is calculated the power P_L to be handled by the shunt stabiliser. This consists of three components: P_v, the power to compensate for the variations in the mains voltage; P_c, the power to compensate for component tolerance; and P_b, the beam current compensation.

With the above values in hand, it is possible to calculate the values of components in the h.t. line and the shunt stabilising circuit.

From Appendix 1, the following equations may be obtained. Using Eq. 3 or Eq. 4,

$$P_v = zP. \qquad \ldots(10)$$

Using Eq. 8 (P_t—component tolerance)

$$P_c = cP_t. \qquad \ldots(11)$$

Using Eq. 8 (P_b—e.h.t. power)

$$P_b = cP_b \qquad \ldots(12)$$

P_a = power taken by d.c. component when audio changes from 0 to full rated power ...(13)

$$P_L = P+zP+cP \qquad \ldots(14)$$

H.T. CIRCUIT

The h.t. current supplied to the line output transformer is

$$I_{ht} = \frac{P_L}{V_3}$$

The h.t. current taken by video and line driver stages is

$$I_{video} = \frac{P_{video}}{V_6}$$

Calculations indicate that h.t. rectifier and 2·7Ω series resistor can be approximated by an equivalent resistor of 63Ω when the input voltage is taken as $V\sqrt{2}$, where V is the r.m.s. value of nominal mains voltage.

$$V_1 = V\sqrt{2} - 63\left[\frac{P_L}{V_3} + \frac{P_{video}}{V_6}\right] \qquad \ldots(15)$$

and

$$R_1 = \frac{V_1 - V_3}{I_{ht}} - 56. \qquad \ldots(16)$$

The 56Ω in the above equation represents allowances for the flashover protection resistor.

REQUIREMENT OF SHUNT STABILISER CIRCUIT

The operation of the shunt stabiliser circuit may be visualised more easily by considering it in terms of currents instead of power. It will be assumed that a circuit concept as shown in Fig. 1 is used. In this case the shunt stabiliser is given a low internal impedance and the purpose of R_3 is to increase the circuit impedance to the level required by the line output transformer for a correct compromise in respect of raster 'breathing'. Normally the difference between V_4 and V_5 is quite small, and we may say that $V_5 = 0{\cdot}95V_4$.

Remembering that P_L represents power requirements under nominal conditions, including component tolerances, and full output power rating, then the current to be drawn by the shunt stabiliser is:

$\frac{P_v}{V_4}$ with nominal mains voltage

or 0 with mains voltage at −10% (see Eq. 6)

or $2\frac{P_v}{V_4}$ with nominal mains voltage +10%.

The effect of component tolerances is to modulate this current, irrespective of mains voltage, by $\pm P_c/V_4$. When the audio power is turned down, the excess current to be taken by the shunt stabiliser, irrespective of mains voltage is P_a/V_4. Finally, when the beam current is reduced from 300μA to 0, the excess of current, irrespective of mains voltage, is P_b/V_4.

The current, independent of mains voltage variation, is

$$\frac{1}{V_4}(P_a+P_b\pm P_c).$$

Thus the maximum current passing through the shunt stabiliser is as follows;

with −10% mains, $\frac{1}{V_4}(P_a+P_b\pm P_c)$

with nominal mains, $\frac{1}{V_4}(P_v+P_a+P_b\pm P_c)$

with +10% mains, $\frac{1}{V_4}(2P_v+P_a+P_b\pm P_c)$. ...(17)

Since the latter entry is the largest quantity, this represents the maximum current of the shunt stabiliser, and in fact the current of the power control device in this circuit, that is, I_{TR1}.

With reference to the first of the above three values, it should be noted that when the receiver is operating at full beam current and full audio output, $P_a = 0$ and $P_b = 0$, and therefore $I = \pm P_c/V_4$. It is obvious that the circuit cannot provide a negative current. Therefore all currents should be raised by P_c/V_4. However, the coincidence of minimum mains voltage, maximum audio output, maximum beam current and maximum component tolerances should be rare and for this reason the above term is neglected.

The value of resistor R_2 in series with TR_1 (Fig. 1) is given by

$$R_2 = \frac{(0{\cdot}95)V_4}{\frac{1}{V_4}(2P_v+P_a+P_b+P_c)} = \frac{(0{\cdot}95)V_4^2}{2P_v+P_a+P_b+P_c} \qquad \ldots(18)$$

From this, the maximum possible dissipation of TR_1 is thus

$$P_{TR1} \simeq \frac{V_4^2}{4R_2} = \tfrac{1}{4}(2P_v+P_a+P_b+P_c) \qquad \ldots(19)$$

AVERAGE CURRENT OF SCAN RECTIFICATION DIODE

The maximum value of the average current passed by the scan rectification diode is given by

$$I_d = I_{TR1} + \frac{\left(\begin{array}{c}\text{Power taken from scan rect. supply,}\\ \text{less audio, and less shunt stab. power}\end{array}\right)}{V_4} \qquad \ldots(20)$$

Example 1

A monochrome receiver is designed for 20kV operation (I_b = 300μA). For the line output transformer used, the optimum condition in respect of raster 'breathing' is when the h.t. applied to it varies ±3·6% (x = 0·036) for ±10% change in the mains voltage (p = 0·1). The transformer is designed to operate from an h.t. line of 135V. An SEPP field timebase is employed and 2W audio output is required. The low voltage supply is to be 32V.

Power requirements are as follows:

Line timebase	17W
E.H.T.	6W
Field timebase	13W
Audio	3·5W
R.F., i.f., jungle, etc.	3·6W
Stabilised circuit services	1W
Other losses	0·5W

$P = 44{\cdot}6W$

$P_B = 6W$

$P_a = 3W$

Evaluation of constants:

$$k = \frac{\sqrt{2}\times 240}{135} = 2{\cdot}5$$

$$r = \frac{k(1-p)-(1-x)}{1-x} = \frac{0{\cdot}9\times 2{\cdot}5-(1-0{\cdot}036)}{1-0{\cdot}036} = 1{\cdot}334$$

$$z = \frac{k(p-x)}{k(1-p)-(1-x)} = \frac{2{\cdot}5(0{\cdot}1-0{\cdot}036)}{0{\cdot}9\times 2{\cdot}5-(1-0{\cdot}036)} = 0{\cdot}1245$$

$$n = \frac{P}{P-P_B} = \frac{44{\cdot}6}{38{\cdot}6} = 1{\cdot}16$$

$$c = \frac{1}{1+\left[\frac{1}{r}+\frac{1}{n}\right]\frac{x}{z}} = \frac{1}{1+\frac{1}{1{\cdot}334}+\frac{1}{1{\cdot}16}\quad\frac{0{\cdot}036}{0{\cdot}1245}} = 0{\cdot}68$$

Assuming ±20% spread in power requirements due to component tolerances the following values may be derived;

	s	s^2
Line timebase tolerance	3·4W	10·6
Field timebase tolerance	2·6	6·8
R.F., i.f., jungle tolerance	0·72	0·52
Other tolerance	0·3	0·09
		18·01

$$P_t = \sqrt{18} = 4{\cdot}25\text{W}$$

Since these tolerances are independent of each other, they are added statistically, thus:

$$P_v = zP = 0{\cdot}1245 \times 44{\cdot}6 = 5{\cdot}5\text{W}$$
$$P_c = cP_t = 0{\cdot}68 \times 4{\cdot}25 = 2{\cdot}9\text{W}$$
$$P_b = cP_B = 0{\cdot}68 \times 6 = 4{\cdot}1\text{W}$$
$$P_L = P + zP + cP_t = 44{\cdot}6 + 5{\cdot}5 + 2{\cdot}9 = 53\text{W}.$$

H.T. current supplied to the line output transformer is then given by

$$I_{ht} = \frac{53}{135} = 0{\cdot}393\text{A}.$$

Taking h.t. line to the video and line driver as $V_6 = 150$V, the corresponding current is 10·5/150 = 0·07A.

Then

$$V_1 = \sqrt{2}V - 63(0{\cdot}393 + 0{\cdot}07)$$

$$V_1 = 340 - 29{\cdot}5 = 310\text{V}$$

and

$$R_1 = \frac{310 - 135}{0{\cdot}39} - 56 = 394\text{V}.$$

The value of I_{TR1} is then calculated as follows:

$$I_{TR1} = \frac{1}{V_4}(2P_v + P_a + P_b + P_c)$$

$$= \frac{11 + 2{\cdot}8 + 4{\cdot}1 + 2{\cdot}9}{32} = \frac{20{\cdot}8}{32} = 0{\cdot}65\text{A}$$

$$R_2 = \frac{(0{\cdot}95 \times 32) - 1}{0{\cdot}65} = \frac{29{\cdot}4}{0{\cdot}65} = 45\Omega.$$

Allowing for tolerances, a 42Ω resistor may be used, the dissipation being:

$$P_{TR1} = \frac{V^2}{4 \times 42} = \frac{32 \times 32}{168} = 6{\cdot}1\text{W}.$$

The power taken from the low voltage rail (less audio and shunt stabilisation supply) is:

Field timebase	13W
Audio (standing power)	0·5W
R.F., i.f., jungle and losses	5·1W
	18·6W

This corresponds to a current of 0·58A and therefore the total average current through the scan rectification diode is

$$I_d = 0{\cdot}58 + 0{\cdot}65 = 1{\cdot}23\text{A}.$$

The values provided by this example are used now to illustrate changes of current taken by the shunt stabilising transistor as a function of mains voltage and load.

It was stated on page 165 that with nominal mains voltage, full outputs and nominal components, the current taken by the shunt stabilising circuit should be

$$\frac{P_v}{V_4} = \frac{5{\cdot}5}{32} = 0{\cdot}172\text{A}.$$

When the mains voltage is changed by −10% this current drops to zero and with mains voltage at +10%, the value of current is doubled. This situation is shown in Fig. 15 as a dotted line.

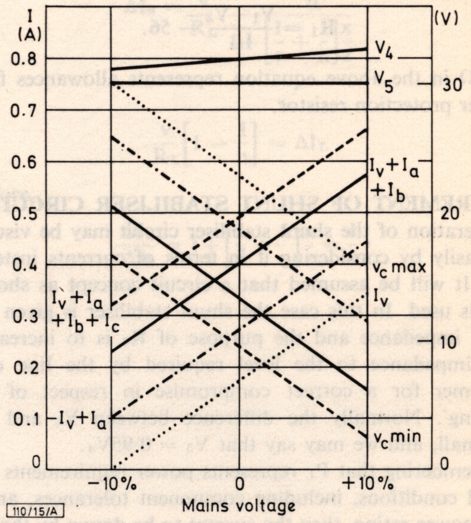

Fig. 15—Calculated change in Ic and Vc

A reduction in audio output to zero produces an increase of current by P_a/V_4 (actually $P_a/0{\cdot}95V = 0{\cdot}1$A) irrespective of mains voltage variations as shown by the dot-dash line.

If the picture tube is cut off there is a further increase of current by $P_b/V_4 = 4{\cdot}1/135 = 0{\cdot}128$A (full line). And finally the component tolerances increase the current by $P_c/V_4 = 2{\cdot}9/32 = 0{\cdot}09$A (dashed line).

Resulting collector voltages for $R_2 = 42$ are also shown in Fig. 15 using graph markings corresponding to their currents.

Example 2

Receiver as in *Example 1*, but operated at 18kV and with class B field output stage (26V low voltage rail).

Power requirements:

Line timebase	15·3W
E.H.T.	5·4W
Field timebase	5·5W
Audio	3·5W
R.F., i.f., jungle	3·0W
Stabilisation circuit services	1·0W
Other losses	0·5W

$$P = 34{\cdot}2\text{W}$$
$$P_B = 5{\cdot}4\text{W}$$
$$P_a + 3\ \text{W}$$

$z = 0{\cdot}1245$
$r = 1{\cdot}334$

$$n = \frac{P}{P - P_B} = \frac{34{\cdot}2}{29{\cdot}8} = 1{\cdot}148$$

$$c = \frac{1}{1 + \left(\frac{1}{r} + \frac{1}{n}\right)\frac{x}{2}} = \frac{1}{1 + (0{\cdot}748 + 0{\cdot}872) \times 0{\cdot}289} = 0{\cdot}682$$

Spreads due to components

	s	s^2
Line timebase	3·06	9·35
Field timebase	1·1	1·21
R.F., i.f., jungle	0·6	0·36
Other	0·3	0·09
		11·01

$$P_t = \sqrt{11} = 3{\cdot}32\text{W}$$
$$P_v = zP = 0{\cdot}1245 \times 34{\cdot}2 = 4{\cdot}25\text{W}$$
$$P_c = cP_t = 0{\cdot}682 \times 3{\cdot}32 = 2{\cdot}26\text{W}$$
$$P_b = cP_B = 0{\cdot}682 \times 5{\cdot}4 = 3{\cdot}68\text{W}$$
$$P_L = P + zP + cP_t = 40{\cdot}71\text{W}$$
$$I_{ht} = \frac{40{\cdot}71}{135} = 0{\cdot}301$$

$$I_{video} = \frac{10{\cdot}5}{150} = 0{\cdot}07\text{A}$$

$$V_1 = 340 - (63 \times 0{\cdot}371) = 317{\cdot}6\text{V}$$

$$R_1 = \frac{317{\cdot}6 - 135}{0{\cdot}301} - 56 = 547$$

$$I_{TR1} = \frac{2P_v + P_a + P_b + P_c}{V_4} = 0{\cdot}672\text{A}$$

$$R_2 = \frac{(0{\cdot}95 \times 26) - 1}{0{\cdot}672} = 35{\cdot}2$$

Allowing for 5% tolerance, $R_2 = 33\Omega$.

$$P_{TR1} = \frac{V_4^2}{4 \times 33} = 5{\cdot}1\text{W}.$$

Power taken from the low voltage rail (less audio and shunt stabilised supply):

Field timebase	5·5W
Audio (standing power)	0·5W
R.F., i.f., jungle, losses	4·5W
	10·5W

Corresponding current = 0·403A.
Total average current of scan rectification diode

$$I_d = 0{\cdot}672 + 0{\cdot}403 = 1{\cdot}075\text{A}.$$

1920—1970

MULLARD TECHNICAL COMMUNICATIONS is the company's major applications journal. It was founded in 1952, and its first and almost only *Editorial* said that its purpose was to make Mullard applications information (until then circulated in a rather limited way) available to 'a still wider circle of professional engineers, and also to technical colleges, research organisations, and technical libraries'. After eighteen years, 107 issues, two formats (a record low number for an electronics journal?), one printer, and three editors, its policy, its content, and its classes of readership remain unchanged. One might only replace, or augment, the somewhat grandiose phrase 'professional engineers' by 'practical designers in virtually every industry throughout the world'.

MTC goes back (or has come forward) a mere eighteen years. Its present issue almost coincides with the fiftieth anniversary of the Mullard company—an occasion that can be appropriately marked by a glance at the earlier days of Mullard applications activity.

It was realised, early in the company's history, that to sell a technological product one must also sell performance and provide performance information. The technological product was the thermionic valve, for in the 1920s and early 1930s we made valves and not much else. Electronics *was* valves.

Testing meant simple d.c. measurements based on published I_a/V_g and I_a/V_a curves, static checks on 'm, Z, k_1' (to become 'g, r, s' and then 'μ, r_a, g_m'), microphony tests, insulation tests, and a reverse grid current test (we called it 'backlash') to reveal 'softness'. The first microphony test used a lead pellet dropped three times by hand (a tedious job) down a tube onto the mounting board of the test gear. The highest flick of the needle determined the fate of the valve. Things were as simple as that.

But all this was not enough. A valve might pass every test and still behave with unpredictable poorness under dynamic conditions at audio or radio frequencies. Capacitive effects, damping, and unknowns that were more elusive and less easily dealt with than HAC, LAC, LS, LE, SOFT, MIC or any other valve malady, had to be considered. Further, it was obviously desirable that customers should use valves in ways that would not only be non-destructive but would allow the valve to give the best performance of which it was capable.

To find out how valves would really behave, to promote the right kind of use by customers, and to discover the specifications of tomorrow's valves, the obvious course was to design practical circuits, to study valve performance in these circuits, to devise better circuits, to design more advanced valves, and to recommend circuit configurations and model operating conditions to the customer—who might be a radio receiver manufacturer, a ham, a government department buying large-glass or silica transmitting valves, or that early D-I-Y man the kitchen-table constructor of a modest two-valver.

Valve testing, about 1930

The earliest discoverable applications advice (from 1924, the year of Wembley and its broadcast just-hearable royal opening) included this warning:

> If the makers say that the filament voltage should be 3·5, do not use six volts because by doing so you get rather louder signals.

Happy days! And, incidentally, we hope that MTC prints few sentences as ambiguous as that.

Of course, in the early thirties electronics meant mainly 'wireless'—both ends of it: the thermionic triode; the screen-grid tetrode with its capacitance-reducing but mechanically troublesome top-cap anode connection (threads stripped, cement powdered); the early a.f. pentodes; the rectifiers that, when they first had cathodes, punctured alarmingly on life test; unwieldly passives and wound components; the protracted death of 'jars'; the dart-board loudspeakers; the brass terminals and black Ebonite; and the polished mahogany boxes.

The applications laboratory was correspondingly simple. It was housed, around 1931, in the attics of a garage just down the road from the valve factory in Nightingale Lane. The laboratory, with a staff of two, was part of the seven-man five-boy Technical Department (known with awed affection as 'TD'), which handled everything from valve development through unsophisticated quality control to seemingly endless life tests, and applications and customer problems. The present editor of MTC remembers selecting PM4 triodes for a very special customer—the 1935 Everest expedition; and has, ever since, hoped that that expedition's comparative non-success owed nothing to faulty selection, to untypically poor customer service.

The growth of electronics since the thirties, in volume and scope and complexity and power and frequency coverage, needs no retelling here. It is reflected in the great expansion of the Mullard applications activity—now carried forward by a hundred engineers in the Central Application Laboratory, who have vastly more difficult tasks than the selection of filamentary triodes for running with Megger-generated h.t. And it is reflected in the pages of Mullard books, Mullard booklets and, of course, MULLARD TECHNICAL COMMUNICATIONS.

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